VOLUME 24 APRIL, 1936 NUMBER 4

PROCEEDINGS
of
The Institute of Radio Engineers

Eleventh Annual Convention
Cleveland, Ohio
May 11, 12, and 13, 1936

Application Blank for Associate Membership on Page XV

Institute of Radio Engineers Forthcoming Meetings

ELEVENTH ANNUAL CONVENTION
May 11, 12, and 13, 1936
Cleveland, Ohio

JOINT MEETING
American Section, International Scientific
Radio Union and Institute of Radio Engineers
WASHINGTON, D. C.
May 1, 1936

BOSTON SECTION
April 24, 1936

CLEVELAND SECTION
April 26, 1936

EMPORIUM SECTION
April 24, 1936

NEW YORK MEETING
April 1, 1936

PHILADELPHIA SECTION
April 2, 1936
May 7, 1936

WASHINGTON SECTION
April 13, 1936

INSTITUTE NEWS AND RADIO NOTES

March Meeting of the Board of Directors

The regular monthly meeting of the Board of Directors was held in the Institute office on March 4, 1936. Those present were Alan Hazeltine, president; Melville Eastham, treasurer; E. H. Armstrong, Arthur Batcheller, Virgil M. Graham, L. C. F. Horle, C. M. Jansky, Jr., H. M. Turner, William Wilson, and H. P. Westman, secretary.

E. F. Carter, W. E. Holland, and I. G. Maloff were transferred to Fellow grade, and H. A. Chinn, G. L. Haller, H. A. Robinson, and H. R. Skifter were transferred to Member. J. I. Cornell, H. F. Humphreys, J. G. Robb, and F. L. Smith were admitted to Member grade. Seventy-four applications for Associate membership and eighteen for Student grade were approved.

Some additional committee appointments were made and the complete list of those serving on Institute committees is given elsewhere in this issue.

Joint IRE-URSI Meeting

A joint meeting of the American Section of the International Scientific Radio Union and the Institute of Radio Engineers will be held on May 1, 1936. There will be two sessions at the building of the National Academy of Sciences, 2101 Constitution Avenue, Washington, D.C., beginning at 10 A.M. and 2 P.M. Papers will be limited to fifteen minutes each to allow time for discussion. The preliminary program given below had been arranged at the time of going to press.

It is expected that reduced railroad rates will be available to members of the IRE and URSI from April 20 to May 2 and that tickets so purchased will be good for thirty days. If the special rates are made available, the certificates required for validation may be secured by application about April 1, to S. S. Kirby, Technical Secretary, American Section URSI, National Bureau of Standards, Washington, D.C.

PRELIMINARY PROGRAM

"Solar Radio Disturbances," by J. H. Dellinger, National Bureau of Standards.

"Polarization of Radio Waves from the Ionosphere near the Geomagnetic Equator," by L. V. Berkner and H. W. Wells, Carnegie Institution of Washington.

"Multifrequency Ionosphere Recording," by T. R. Gilliland, National Bureau of Standards.

"Anomalies in the Magnetoionic Difference Frequency of the Ionosphere," by J. P. Schafer and W. M. Goodall, Bell Telephone Laboratories.

"Recent Correlations between the Ionosphere and High-Frequency Radio Transmission," by N. Smith, S. S. Kirby, and T. R. Gilliland, National Bureau of Standards.

"Further Observations of Ultra-High-Frequency Signals over Long Indirect Paths," by R. A. Hull, American Radio Relay League.

"The Reliability of Short-Wave Radiotelephone Circuits," by R. K. Potter and A. C. Peterson, Bell Telephone Laboratories.

"Ultra-High-Frequency Transmission between RCA building and Empire State building in New York City," by P. S. Carter and G. S. Wickizer, R.C.A. Communications.

"Organization of Activities for Survey Measurement and Elimination of Man-Made Interference," by L. C. F. Horle, consulting engineer.

"Application of Fourier's Theorem to Directional Antenna Design," by I. Wolff, RCA Manufacturing Company.

"Transmission of Electromagnetic Waves in Hollow Tubes," by W. L. Barrow, Massachusetts Institute of Technology.

"Radiation from an Antenna over a Conducting Earth," by W. W. Hansen and J. G. Beckerley, Stanford University.

"Development of an Improved Apparatus for the Measurement of Dielectric Constants by the Heterodyne Beat Method," by R. T. Gabler, Carnegie Institute of Technology.

"Applied Electron Optics," V. K. Zworykin, RCA Manufacturing Company.

"Thermionic Tube Technique for the Measurements of Dielectric Constants of Gases at Radio Frequencies," by L. G. Hector, University of Buffalo.

"Vacuum Tube Voltmeters for use at Ultra-High Frequencies," by L. S. Nergaard, RCA Manufacturing Company.

"Cathode-Grid Bias in Tube Operation," by I. E. Mouromtseff, Westinghouse Electric and Manufacturing Company.

"Unicontrol Radio Receiver for Ultra-High Frequencies Using Concentric Lines as Interstage Couplers," by F. W. Dunmore, National Bureau of Standards.

Intended Substitution of the Practical Absolute System of Electrical Units for the Existing International System

The following information was approved for general publication by the International Committee of Weights and Measures at its meeting in October, 1935, at Sevres, France.

(1) In accordance with the authority and responsibility placed upon it by the General Conference of Weights and Measures in 1933, the International Committee of Weights and Measures has decided that the actual substitution of the absolute system of electrical units for the international system shall take place on the first of January, 1940.

(2) In collaboration with the national physical laboratories, the Committee is actively engaged in establishing the ratios between the international units and the corresponding practical absolute units.

(3) The Committee directs attention to the fact that it is not at all necessary for any existing electrical standard to be altered or modified with a view to making its actual value conform with the new units. For the majority of engineering applications the old values of the international standards will be sufficiently close to the new for no change, even of a numerical nature, to be required. If for any special reason a higher precision is necessary, numerical corrections can always be applied.

(4) The following table gives a provisional list of the ratios of the international units to the corresponding practical absolute units, taken to the fourth decimal place. Since differences affecting the fifth decimal place exist between the standards of the international units held by the various national laboratories and also because all the laboratories which have undertaken determinations of the values of their standards in absolute measure have not yet obtained final results, the committee does not consider it desirable for the present to seek a higher precision. At the same time it hopes that it will be possible to extend the table of these ratios with a close approximation to the fifth decimal place well before the date fixed for the actual substitution of the practical absolute system for the international system.

1 Ampere	International	=0.999 9	Ampere Absolute
1 Coulomb	"	=0.999 9	Coulomb "
1 Ohm	"	=1.000 5	Ohm "
1 Volt	"	=1.000 4	Volt "
1 Henry	"	=1.000 5	Henry "
1 Farad	"	=0.999 5	Farad "
1 Weber	"	=1.000 4	Weber "
1 Watt	"	=1.000 3	Watt "
1 Joule	"	=1.000 3	Joule "

Radio Library Presented to Canada

Noting that technical papers prepared by Canadian engineers on communications subjects disclosed lack of a wide range of references, Donald McNicol, a Fellow and Past President of the Institute, has presented a large and valuable library of published books and technical papers to Queens University. These works include historical and modern items on telegraphy, submarine cables, telephony, radio, television, and talking picture engineering, a total of more than one thousand items including books and papers. It contains numerous original documents of historical interest and of which there are few, if any, duplications. The collection will be housed in the Douglas Library at Queens University, Kingston, Canada, for reference use, and will be known as the McNicol Library on Communication.

Committee work

Membership Committee

The Membership Committee met in the Institute office on March 4. Those present were F. W. Cunningham, chairman; H. A. Chinn, J. M. Clayton. I. S. Coggeshall, E. D. Cook, H. C. Gawler, H. C. Humphrey, E. W. Schafer, Leslie Woods, and H. P. Westman, secretary. A major portion of the time was devoted to a discussion of higher grade memberships.

Technical Committee on Radio Receivers

The Subcommittee on Test Procedures of the Technical Committee on Radio Receivers met in the Institute office on March 5. D. E. Foster chairman; L. F. Curtis, E. T. Dickey, H. A. Wheeler, and H. P. Westman, secretary, were present. H. O. Peterson, a member of the Technical Committee on Radio Receivers, was present as a visitor and Lloyd Espenschied and C. E. Pfautz attended in behalf of a committee which is preparing material for use at a forthcoming C.C.I.R. conference.

The subcommittee completed its consideration of the existing report and it is anticipated that its final report to the technical committee can now be made available.

It discussed with the representatives of the C.C.I.R. preparatory committee, the preparation of a special report on the testing of broadcast receivers for use at a forthcoming international conference.

Institute Meetings

Atlanta Section

The annual meeting of the Atlanta Section was held on January 16 at the Atlanta Athletic Club. I. H. Gerks, chairman, presided and there were eighteen members and guests present, eight of whom attended the informal dinner which preceded the meeting. In the election of officers, Professor Gerks was re-elected chairman, P. C. Bangs of the Acoustic Equipment Company was named vice chairman, and N. B. Fowler of the American Telephone and Telegraph Company was designated the secretary-treasurer.

Professor Gerks presented a paper on "Stabilized Feed-Back Amfiers" which was introduced with a brief outline of feedback and the basic theory of amplification. Controlled feed-back amplifiers were then discussed and it was shown that when the feed-back voltage is of proper phase and magnitude, oscillation occurs. Use of feed-back voltage may be made to reduce noise, harmonic distortion, and hum in amplifiers. The paper was discussed by Messrs. Bangs, Fowler, and Reid.

Boston Section

R. G. Porter, secretary-treasurer, presided at the December 20 meeting of the Boston Section which was held at Harvard University. There were ninety present and seven attended the informal dinner which preceded the meeting.

A paper by E. L. Chaffee, professor of physics of Harvard University was on "The Operating Characteristics of Power Tubes as Determined by Experimental Tests at Low Frequency." It was pointed out that operating characteristics of power tubes have hitherto been determined either by operating the tube under test at a radio frequency and making necessary measurements or by laborious calculations assuming an ideal path of operation or using static characteristics as a starting point. Dr. Chaffee presented a method in which measurements are made at low frequency with magntitudes of grid and plate voltages of the order employed in normal operation at a radio frequency. Test results were given by contours of constant power output, efficiency, tube loss, etc., on the e_p-e_g plane. These contours reveal best operating conditions for class B or C amplifiers or for operation as a self-excited oscillator. A method was presented for calculating contour charts from static characteristic curves and was based on a simplified method of harmonic analysis. The paper was discussed by Messrs. Barrow, Field, and Thiessen.

Buffalo-Niagara Section

"Photoradio Analogs" was the subject of a paper by Austin Armer of the Magnavox Company given at the February 12 meeting of the Buffalo-Niagara Section, which was held at the University of Buffalo. L. E. Hayslett, chairman, presided and there were eighty present.

The speaker showed graphs illustrating the performance of photographic emulsions and light filters with respect to frequency, speed, range, and intensity. These curves were compared with similar graphs of the characteristics of radio tubes and circuits. He discussed some existing phases connected with motion picture film having photographic sound track, and color, astronomical and other types of photography.

Chicago Section

H. S. Vance, chairman, presided at the February 21 meeting of the Chicago Section which was held in the RCA Institutes Auditorium. The meeting was attended by 160, of whom twenty-six were present at the dinner which preceded it.

A paper on "NBC Directional and Nondirectional Antenna Developments and Notes on Factors Affecting Broadcast Station Cover-

age" was presented by R. F. Guy and W. S. Duttera, radio engineers of the National Broadcasting Company.

Mr. Guy presented the first part of the paper in which he discussed factors determining the coverage of broadcast transmissions and their relative importance. An antenna of uniform cross section with a lumped capacitance and tuning system at the top which permits the electrical length to be varied over wide limits was described. Measurements and fading records comparing this type with older designs showed definite advantages in favor of the newer form. Recordings from WPTF Raleigh, North Carolina, as far away as to the Rocky Mountains were shown as well as measurements taken in airplanes to indicate the vertical field distribution.

Mr. Duttera covered directive systems using two similar radiators and analyzed the effects of varying the phase relation between similar currents in two radiators and the spacing of them. A graphical method of calculating radiation from directive antenna systems employing radiators having any vertical characteristic was given. The radiation at high angles and along the ground were compared with various operating conditions and antennas having electrical lengths of ninety and 190 degrees. It was shown that high angle radiation is increased in certain types of directive systems and that consideration of this is of less importance where radiators inherently give small high angle radiation. A comparision of directive and nondirective transmissions from WPTF indicated a gain of one hundred and twenty-five to one in voltage when employing a cardiod as compared with a circular pattern.

Cincinnati Section

On February 6 the Cincinnati Section participated in the first joint meeting of the technical and scientific societies of Cincinnati which is comprised of ten local groups. The meeting was held at the Music Hall in Cincinnati and L. L. Bosch, chairman of the Council of these affiliated societies, presided. The attendance was 3000 and there were seventy at the informal dinner which preceded the meeting.

"Man's Farthest Aloft" was the subject of a paper by A. W. Stevens Captain, U. S. Army Air Corps. He described stratospheric attempts made previous to his record-breaking flight of November 11, 1935. He then covered the preparations, take-off, flight and landing of the *Explorer II* and showed still and motion pictures taken from the balloon, the ground, and airplanes.

Cleveland Section

The January 23 meeting of the Cleveland Section was held at the Case School of Applied Science and presided over by R. M. Pierce,

chairman. There were sixty-three members and guests present and eighteen at the informal dinner which preceded the meeting.

A paper on "Getting the Most out of the Cathode-Ray Tube" was presented by J. M. Stinchfield of the Research and Development Laboratory of RCA Radiotron. Data were presented on the operating characteristics of cathode-ray tubes, sweep circuits, and input circuits for operating tubes. This was followed by a demonstration showing the effect of the change in capacitance and reistance of these sweep circuits and of other controlling factors.

H. P. Westman, national secretary, spoke briefly of convention plans.

Emporium Section

The Emporium Section met on February 27 at the American Legion Club Rooms. H. A. Ehlers, vice chairman, presided and there were forty-seven present.

L. B. Arguimbau of the General Radio Company presented a paper on "Design and Application of Laboratory Equipment." He traced the development of several widely used laboratory instruments such as the wave analyzer and audio oscillators and described the functions and uses of such equipment. Some new concepts in measuring technique were presented. The paper was discussed by Messrs. Graham, Keen, Kievit, Mahaffey, Overmier, and West.

Detroit Section

E. C. Denstaedt, chairman, presided at the February 28 meeting of the Detroit Section held at the Hotel Statler. There were 225 present and twenty-four at the dinner which preceded the meeting.

A paper on "Modern Adaptations of Electron Tubes in Industry" was presented by R. A. Powers, chief physicist of Electronic Control Corporation. He outlined briefly the history of photoelectric equipment pointing out that a selenium cell for counting objects was described several years before the invention of the electric lamp, the drawings of the device showing a kerosene lamp as the light source. The use of photoelectric devices and methods of causing them to respond to differences in temperature, color, or size was described. A device which is capable of differentiating between temperatures separated by only two degrees at a temperature of the order of 1650 degrees Fahrenheit was demonstrated. It is employed in the hardening of the ends of automobile valve stems. The stem is clamped between two electrodes and is heated by an electric current to proper hardening temperature in approximately half a second at which time the photocell opens the heater circuit and initiates the operation which quenches the metal. The device may be used for determining the temperature of steel billets inside a furnace.

A device which responds to the capacitance of objects entering the field of a small antenna controls the operation of automatic spray painters and permits them to function only when an object to be sprayed is located in front of them. The paper was concluded with a discussion of service difficulties in this field of engineering.

New York Meeting

A meeting of the Institute was held at the Engineering Societies Building in New York City on March fourth and was presided over by President Hazeltine. Six hundred and fifty members and guests were present.

A paper on "The Application of Multipactors to Radio-Frequency Amplifiers and Oscillators" was presented by P. T. Farnsworth, vice president of Farnsworth Television, Inc. He presented as a brief review the history and the present scope of the electron multiplier art. The conditions necessary for electron multiplication were listed and modes of electron oscillation discussed. The requirements for energy transfer between an oscillating electron cloud and a radio-frequency field were pointed out and methods of controlling the amount of multiplication were considered. The construction and theory of a simple type oscillator were discussed and several tubes exhibited and described. Two practical methods for crystal controlling multipactor oscillators were described as was the use of the tube as a radio-frequency amplifier. A simple modulated oscillator of low power was demonstrated as was a two-tube multipactor oscillator-amplifier giving an output of 1000 watts. A number of those present entered into the discussion of the paper.

Philadelphia Section

Knox McIlwain, chairman, presided at the February 6 meeting of the Philadelphia Section which was held at the Engineers Club and attended by 300. There were seventeen at the dinner.

Two papers were presented, and the first by C. B. Foos of the tube engineering department of the General Electric Company was on "High Voltage Mercury-Pool Tube Rectifiers." It covered the ignitron type of rectifier in which conduction is established at the desired point in each cycle by passing a short pulse of current through an electrode of special high resistive material which dips into the mercury pool. The output voltage is controlled by adjusting the point in the cycle where conduction is started and these tubes have been successfully used for high voltage, high power rectifiers. Their advantages over hot cathode tubes lies mainly in their lack of emission and peak current limitations. They are superior to mercury-arc tanks in that they are single sealed-

off half-wave units. A trial installation at the South Schenectady transmitting station was described.

The second paper was on "Electrical Measurements at Ultra-High Frequencies" and L. S. Nergaard of the research and development laboratory of RCA Radiotron presented it. In it Dr. Nergaard pointed out that methods of measuring power, current, and voltage at ordinary radio frequencies have failed to give reliable results in the ultra-high-frequency portion of the spectrum. His contribution to the development of a new technique and new devices for ultra-high-frequency measurements include a signal generator covering from twenty to 200 centimeters, thermo-couples for power and current measurements, and very small diodes for voltage determinations. Their limits and uses were discussed.

Rochester Section

A paper on "Photoradio Analogs" was presented by Austin Armer of the Magnavox Company at the February 13 meeting of the Rochester Section which was held at the Sagamore Hotel. There were fifty members and guests present and nine attended the dinner which preceded the meeting. This paper was identical to that given at the Buffalo-Niagara Section meeting and is reported elsewhere in this issue. The paper was discussed by Messrs. Henderson, Karker, and Schoen.

Seattle Section

On January 31 a meeting of the Seattle Section was held at the University of Washington and presided over by E. D. Scott, chairman, Thirty-one were present.

"A First-Hand View of Radio Development Abroad" was presented by Morris Leviten of the Wedel Company. It was based on tours of Europe, Asia, and Africa made by the author during the past two summers. In his opinion, the development of radio in this country is in advance of the rest of the world although some countries are making great strides in particular fields. The U.S.S.R. is making great progress in developing radio therapy, military radio, and electronic equipment. France is making progress in microwave transmission and with modulated light transmission. Holland has very powerful high-frequency directional transmitters for communication with her colonies, and England likewise employs similar equipment to reach the various parts of her empire. As part of the general discussion, a number of questions regarding military radio equipment of foreign nations were answered.

For some years past, it has been the practice of the Seattle Section to devote its June meeting to the presentation of papers by advanced

students of electrical engineering of the University of Washington. It was agreed that the acceptance of these papers should be placed on a competitive basis and rules governing their grading were adopted. Prizes will be awarded for the best papers submitted.

The February meeting of the Seattle Section held at the University of Washington with E. D. Scott presiding was attended by forty-five.

A paper on "Notes on the Use of Rectifiers in Radio Equipment" was presented by H. J. Price, chief engineer of KXA. In it he discussed the design and use of single and polyphase rectifier circuits. He described briefly high vacuum, mercury-vapor, and copper-oxide rectifiers. The paper was discussed by Messrs. Wallace, Woodyard and others.

Toronto Section

L. M. Price, chairman, presided at the February 10 meeting of the Toronto Section held at the University of Toronto. There were 130 members and guests present and fourteen attended the dinner.

A paper on "Police Radio" was presented by B. deF. Bayley, assistant professor of electrical engineering at the University of Toronto. In it he reviewed the history of police radio from 1916 to date and indicated the wide use of automobiles as an important factor in the necessity for police radio installations. He pointed out that a signal level of one millivolt per meter for business districts, one-fifth millivolt for residential sections, and one-twentieth millivolt for rural areas had been found acceptable and sufficient to override normal noise found in those areas. High fidelity of reproduction is not necessary for this service and may be undesirable from the angle of noise. Receivers need not have a sensitivity in excess of fifty millivolts per meter but the selectivity must be relatively high as channels are eight kilocycles wide. With an average time of responding to calls of about two minutes, police radio cannot fail to have a marked influence on both the prevention and correction of crime. It increases the efficiency of existing personnel by saving from five to thirty minutes on each call.

Washington Section

At the February 10 meeting of the Washington Section, which was held in the Potomac Electric Power Company auditorium, the attendance was fifty. Twenty-one were present at the informal dinner which preceded the meeting. E. M. Webster, chairman of the Papers Committee, presided and a paper on "Theoretical Considerations of Directive Arrays" was presented by G. H. Brown of the engineering department of the research division of RCA Manufacturing Company.

Proceedings of the Institute of Radio Engineers
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TECHNICAL PAPERS

AUTOMATIC COMPENSATION FOR CLASS B BIAS AND PLATE VOLTAGE REGULATION*

By

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Summary—*A method of automatically compensating for the effects of bias and plate-voltage regulation in class B stages is outlined. It obviates the necessity of using heavy current bleeders where rectifier bias supply is used, and is shown to be more desirable than battery bias. A typical installation using low power tubes is described, but the principle may be applied to an installation of any desired power. It is shown further that such a system imposes less severe requirements on the plate supply as regards regulation. The only sacrifice entailed is a slight reduction in peak power output due to reduction in supply voltage on modulation peaks.*

Introduction

THE problem of obtaining bias voltage for class B stages or in fact any stages in which there is a variable flow of grid current long has been more or less difficult of solution, particularly where large tubes involving large values of grid current are used. Common practice among broadcast engineers is to use either storage batteries, large rectifiers with high current bleeders, or the equivalent. Any of these systems is at best only an approximate solution. Even if we assume a bias source of perfect regulation, the fact still remains that the plate-voltage source is not infinitely large and thus must have some voltage fluctuation with load variation such as results with class B operation. It is therefore desirable to reduce the bias voltage by the correct amount to compensate for plate-voltage regulation in order to prevent the class B stage from entering or even approaching the cutoff region of the plate-current versus bias-voltage characteristic. Even with zero bias class B tubes the driver transformer resistance permits an accumulation of negative bias during the flow of grid current, also there exist nonlinear plate-current conditions caused by plate-voltage regulation. Correction of both of these conditions can be accomplished by proper application of the compensation system described below. It is possible to provide, with this system, first, a bias source requiring a very much smaller amount of power since the bleeder may

* Decimal classification: R356. Original manuscript received by the Institute, August 7, 1935.

be of the order of 40,000 ohms or higher for audio power of the order of 300 to 400 watts; second, a bias system with essentially perfect regulation; and third, overcompensation, by which plate-voltage regulation can be corrected, thus preventing the distortion now inherent in class B stages caused by shifting near to or past cutoff of plate current as the grid-voltage swing crosses its axis during conditions of low plate voltage brought about by heavy modulation.

Fundamental Circuit

For the moment let us consider the class B stage as a full wave rectifier (Fig. 1) as it is essentially identical as far as the action of accumulating additional bias during periods of grid current is concerned. That is, the grids, when swung positive, draw current exactly as in the

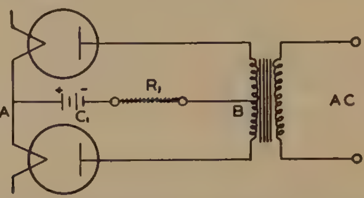

Fig. 1—Basic class B amplifier circuit.

case of a rectifier. This results in developing voltage across the bias circuit resistance R, exactly as a rectifier does across its load. Now if we provide a variable load across AB as in Fig. 2 which is so biased by C_2 that it draws no current from AB until the grid swings positive at G_1G_2 and so adjusted at P_1P_2 as to draw plate current at T_3T_4 equal to the grid current at G_1G_2 there will be no change in the bias voltage at AB. That is, tubes T_3T_4 draw current from circuit AB of an amount exactly equal to the current delivered to circuit AB by grids G_1G_2 acting as rectifiers. Thus the voltage AB will remain constant regardless of the amount of grid current G_1G_2. The only purpose of condenser M is to prevent loss of audio voltage in circuit AB.

These are the conditions for perfect compensation of bias voltage regulation. In practice the fluctuation of bias voltage actually was too small to read on a standard Weston 301 meter where the reading was ninety per cent of full scale.

Overcompensation

The conditions for overcorrection, in which the bias voltage is made to reduce as the plate voltage drops due to plate-voltage power supply regulation, are somewhat more involved for perfect compensation since this correction should start as soon as the class B plate current increases

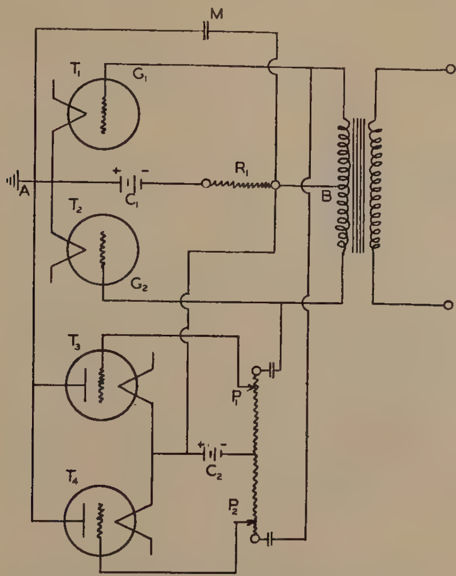

Fig. 2—Fundamental regulator circuit.

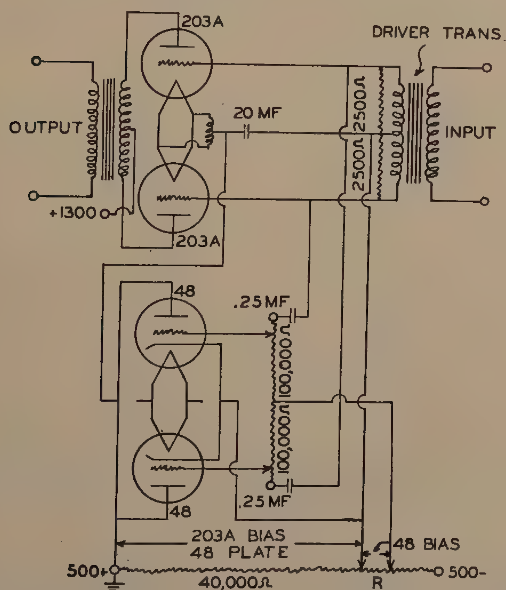

Fig. 3—Typical regulation compensating circuit.

or, in other words, as soon as class B grid excitation starts. Since the compensator tubes are biased past plate current cutoff so as not to provide bias compensation until the class B stage is excited sufficiently

to draw grid current, there can be no overcompensation for plate regulation except in the region of grid current on the class B stage. However, in practice there is generally no appreciable distortion due

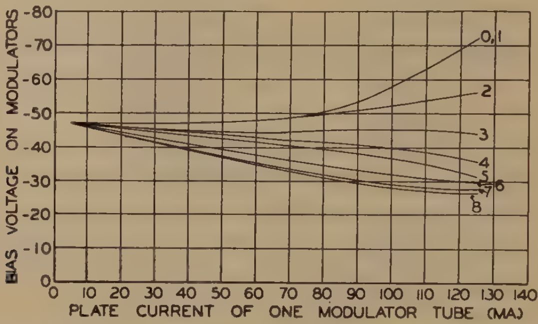

Fig. 4—Bias regulation compensator characteristics. Position 3 used for bias compensation only. Position 5 used for bias and plate supply regulation compensation. Nos. 0–8 represent positions of compensation control.

to plate regulation until the class B stage is fairly well excited, which is generally up to the point that grid current starts to flow, since a slight fall of plate voltage up to this point generally is not sufficient to cut off the class B stage as the grid swing crosses its axis. Beyond the point of

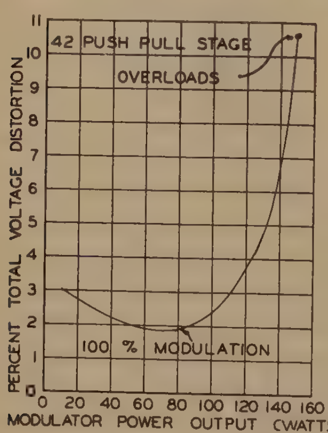

Fig. 5—Modulator output characteristic with bias compensation regulator control set at position 5.

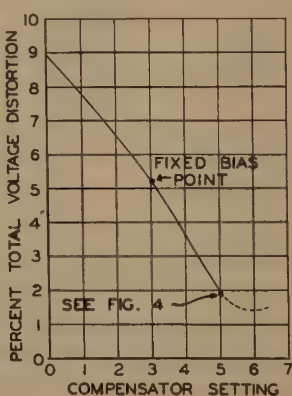

Fig. 6—Harmonic content of modulator output as a function of bias regulation compensator position.

grid current, however, it is desirable to overcompensate the bias circuit, causing the bias to reduce with increased grid swing, thus preventing cutoff of the class B stage and serving not only to increase its out-

put but also to provide a much more linear plate-current—grid-voltage characteristic.

Following is a typical circuit with constants as used at W8BLZ. In this arrangement the adjustments are so set as to provide for bias regulation compensation and partial plate regulation compensation.

Fig. 4 shows the action of the compensator tubes at various settings of the linear potentiometers which control the grid excitation of the type 48 tubes.

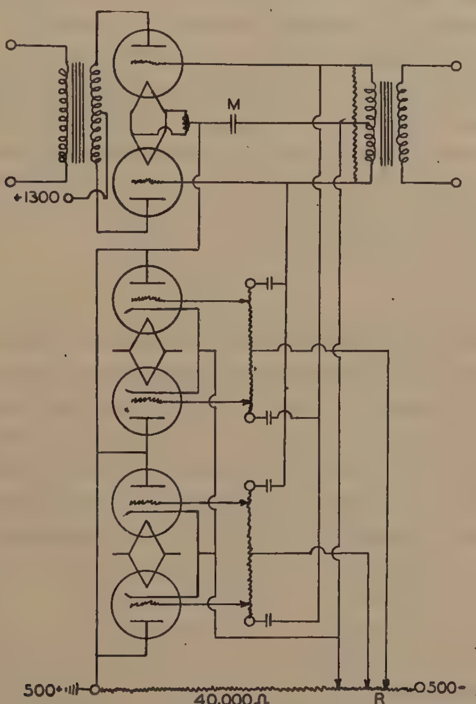

Fig. 7—Fundamental circuit of combination bias and plate voltage supply regulation compensator.

Fig. 5 shows the low distortion resulting from the above arrangement. It is of course understood that the above circuit in no way reduces the power output of the class B stage. The data in Fig. 5 were taken from a setup in which ninety watts output was all that was required for 100 per cent modulation. However, the same class B tubes with larger driver tubes are capable of the same low distortion at several hundred watts output by using the compensator circuit.

It is of further interest to mention that the ninety-watt point on Fig. 5 fell at 5.3 per cent distortion with battery bias and nine per cent distortion with rectifier bias and no compensation. (See Fig. 6.)

For still more perfect compensation of plate regulation it is desirable to make use of two more regulator tubes, the first pair biased past cutoff as described to compensate for bias regulation; the second pair being biased at cutoff to compensate for plate regulation at any value of grid swing instead of in the grid current region only as in the case of the above circuit.

In specifying the correct value for condenser M it is well to remember that this condenser must first serve as audio-frequency return for the grid circuit and therefore must offer negligible impedance at the lowest audio frequency to be transmitted. This involves knowing the impedance of the class B grids at the highest excitation they are to handle. For example, if 203-A tubes are used and are excited to 155 volts above 45 volts bias or 200 volts total, the grid impedance per grid will be E/I, or, for fifty milliamperes grid current, $200/0.050=4000$ ohms. The condenser should therefore not be in excess of 500 ohms reactance at thirty cycles or ten microfarads.

The second function of this condenser is to provide the same time constant in the bias circuit as is present in the plate circuit. That is, the RC product, where R is the bias bleeder resistance (Figs. 3 and 7) and C the condenser, M should essentially be equal to the $R'C'$ product of the plate supply circuit where R' is the resistance of the power supply primary reflected to half the secondary (for full wave rectifier) plus the resistance of half the secondary, plus the filter choke resistance; C' being the total filter capacity. Since condenser M is fixed, it then follows that R must be adjusted so as to make the two RC products equal.

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SCANNING SEQUENCE AND REPETITION RATE OF TELEVISION IMAGES*

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Summary—*This paper considers factors which affect the apparent steadiness of television images: namely, line flicker, flicker of the image as a whole, alternating-current ripple in the deflecting circuits, alternating-current ripple in the video frequency signal, and various kinds of beating of the alternating-current ripple with the various scanning frequencies. It is concluded that an integer ratio between alternating-current ripple frequency and frame frequency is very desirable for progressive scanning and is almost imperative for interlaced scanning. Interlaced scanning with a frame frequency of thirty per second and a field frequency of sixty per second fulfills the requirements in regard to flicker and the relations to alternating-current ripple frequency for a sixty-cycle power source, and offers considerable net gain over other scanning procedures considered. The problems of both odd- and even-line methods of interlacing are discussed and the odd-line method is found preferable.*

Introduction

After considerable experience with the experimental installation previously described,[1] it was concluded that the most objectionable features of the television image were flicker and other unsteadiness.

It is well known that the frequency band required to transmit a television picture is proportional to the product of the picture detail and the frame frequency. (The frame frequency is the number of times per second the picture area is completely scanned.) Since the available frequency band is limited, it is desirable to determine the picture repetition rate which makes the optimum use of the frequency band with regard to picture detail and freedom from flicker. This must be largely a matter of judgment where the psychological aspects are important. As will be seen the decision will be influenced by other factors such as motion picture standards, existing power system frequency, and scanning sequence.

Cathode-Ray Scanning

At present cathode-ray scanning gives the greatest promise for a television system. This involves for example the use of the iconoscope as a pickup device at the transmitter and the kinescope at the receiver.[1] Both devices are similar in that an electrostatically focused

* Decimal classification: R583. Original manuscript received by the Institute, December 18, 1935.

[1] R. D. Kell, A. V. Bedford, and M. A. Trainer, "An experimental television system," Proc. I.R.E., vol. 22, pp. 1246–1265; November, (1934); R. S. Holmes, W. L. Carlson, and W. A. Tolson, "An experimental television system," Proc. I.R.E., vol. 22, pp. 1266–1285; November, (1934).

beam of electrons is deflected vertically by the magnetic field produced by a saw-tooth wave of current (Fig. 1(a)) and horizontally by either a magnetic or electrostatic field produced by a saw-tooth wave of voltage of a much higher frequency (Fig. 1(b)). The action of the two deflecting fields is to cause the light-sensitive screen of the iconoscope or the luminescent screen of the kinescope to be uniformly scanned by the electron beam as shown in Fig. 1(c). In this as well as the following figures, the line frequency is shown as approximately 200 cycles to facilitate illustrating, while in practice it is about 7000 cycles. The return path for each line is shown dotted.

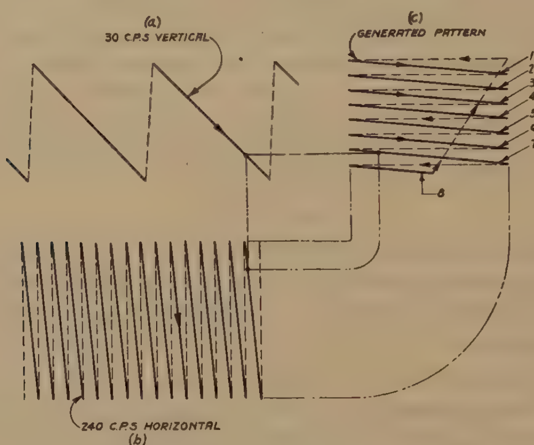


Fig. 1—Generation of a progressive scanning pattern.

Alternating-Current Power Supply Ripple

Alternating-current power-operated receivers contain sixty-cycle and 120-cycle disturbances in the direct voltage supplies which operate the receiver. In sound reception this is called "hum" but in the present paper the term "alternating-current ripple" will be used since the term "hum" implies audibility.

Alternating-current ripple in cathode-ray television shows itself in several ways. When superimposed upon the deflection of the scanning beam it produces the wavy edges as shown in Fig. 2(a), in the case of the horizontal deflection, and causes the nonuniform spacing of the lines of the picture as shown in Fig. 2(b), in the case of the vertical deflection. When the ripple exists in the cathode-ray anode voltage supply, it alters the stiffness of the beam as regards deflection and thereby modulates the deflecting influence of both the deflecting waves. This effect upon the horizontal deflection is shown in Fig. 2(c). It

should be understood that these effects operate not only to distort the shape and density of the scanning pattern, but also cause the misplacement of the details of the picture. The presence of alternating-current ripple in the video frequency amplifier causes the pattern to vary alternately in brightness from top to bottom of the picture. (This effect is not shown in the figures.) Actually all of these effects occur simultaneously though for clearness they have been shown separately.

We now reach a very important point in regard to the psychological effect of alternating-current ripple; i.e., whether the distortion produced (ripple pattern) is stationary or moving with respect to the

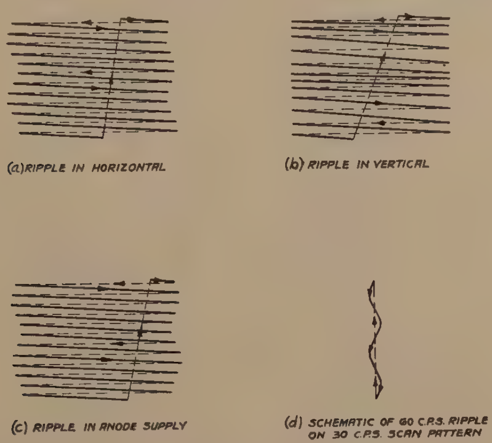
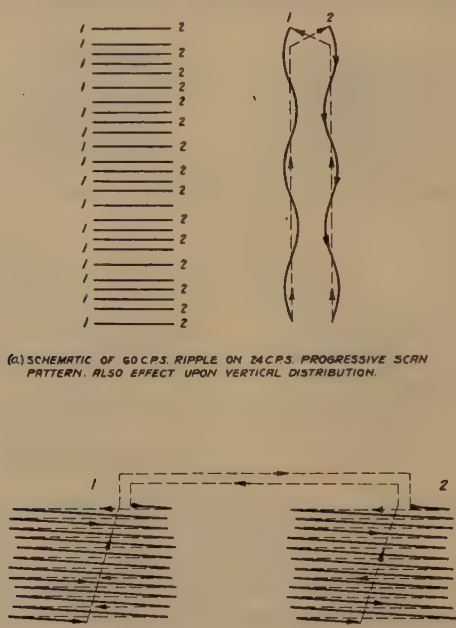


Fig. 2—Effect of sixty-cycle ripple on a thirty-cycle progressive scanning pattern.

scanning pattern. In the case of the thirty-frame per second scanning as shown in Fig. 2, the ripple pattern is stationary and hence is much less objectionable than if it were moving. That the ripple pattern is stationary is evident from a study of Fig. 2(d) and the fact that thirty is a whole number submultiple of sixty. However, in the case of a twenty-four-frame per second picture with sixty-cycle ripple, the alternate frames have distortion of opposite phase (see Fig. 3). This is true since sixty divided by twenty-four is a whole number plus a half, which may be interpreted to mean that the ripple pattern passes over the scanning pattern twelve times per second (since twelve is one half of twenty-four). The figure shows clearly that two vertical scanning cycles are required to make a complete cycle of motion of the beam. This is one of the worst conditions possible, since any element of detail in the picture jumps from one position to another position near by at a frequency of twelve cycles. This causes fatigue of the eye in so far as

the observer is able or attempts to follow the elements of detail in their shifting of position, and a loss of resolution or blurring of the picture in so far as the eye fails to follow the shifting. The presence of sixty-cycle ripple in the picture signal also causes a twelve-cycle flicker of portions of the picture.

In case some picture repetition rate between twenty-four and thirty, say twenty-seven, is chosen, the "ripple pattern" will drift upward across the picture at the rate of about six cycles of ripple per second. At this rate, the eye would be able to follow the elements of picture



(a) SCHEMATIC OF 60 C.P.S. RIPPLE ON 24 C.P.S. PROGRESSIVE SCAN PATTERN. ALSO EFFECT UPON VERTICAL DISTRIBUTION.

(b) ALTERNATE VERTICAL CYCLES FOR 24 C.P.S. PROGRESSIVE SCAN PATTERN WITH 60 C.P.S. RIPPLE ON THE HORIZONTAL DEFLECTION.

Fig. 3

detail so that no appreciable loss of resolution would be observed but due to the propagation of the "ripple pattern" the picture would given an annoying effect of motion, similar to that experienced when viewing stationary objects submerged in water having waves on its surface. Figs. 4(a) and (b) show schematically some standing ripple patterns for various scanning conditions, while (c), (d), and (e) show some conditions for moving ripple patterns that repeat in two or three vertical scanning cycles.

From the foregoing considerations and from tests, it seems desirable that the picture repetition rate be some integer submultiple of sixty,

such as ten, fifteen, twenty, or thirty. So far as continuity of motion in the picture is concerned, it is probable that fifteen or twenty would be high enough, although the motion picture standard is twenty-four per second. However, a much higher repetition rate is required in order to reduce picture flicker to a satisfactory level. From data previously presented about forty-eight pictures per second would be required even if no allowance is made for increase in picture brightness by future development.[2]

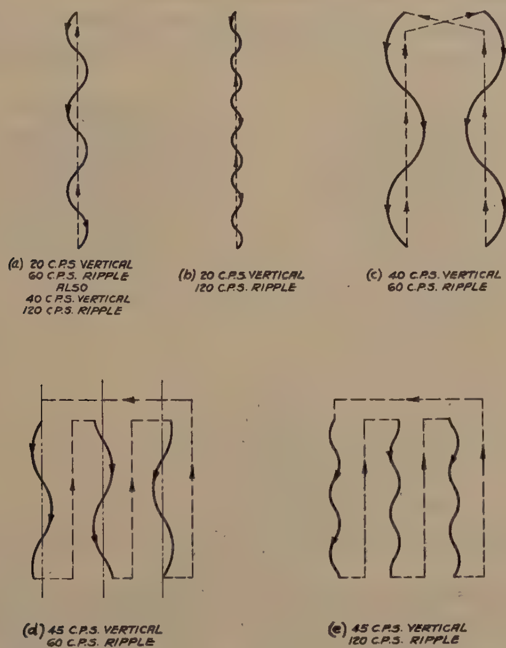


Fig. 4—Schematic representation of alternating-current ripple on various scanning patterns.

Since sixty has no integer submultiple between thirty and sixty, it would seem that the adoption of a repetition rate of sixty per second is required for operation from a sixty-cycle power source. Increasing the picture repetition rate and hence the frequency band in the ratio of sixty to fifteen to eliminate flicker may be giving flicker elimination the advantage in the compromise with picture detail. Yet numerous tests have shown that thirty frames per second is distinctly unsatisfactory in regard to flicker.

[2] E. W. Engstrom, "A study of television image characteristics—Part II," Proc. I.R.E., vol. 23, pp. 295–310; April, (1935).

Odd-Line Interlaced Scanning

At least a partial solution of the problem has been provided by interlaced scanning, in which alternate lines are scanned in successive vertical deflection cycles. Such scanning procedure is old in the television art as produced by the scanning disk in which two or more spirals of apertures are used. In the case of two spirals the interlacing is obtained by locating the apertures on the disk radii such that the apertures of one spiral scan the even-numbered lines and the apertures of the second spiral scan the odd-numbered lines.

An interlaced scanning pattern must be obtained in cathode-ray television by electrical methods. The odd-line method, as the name im-

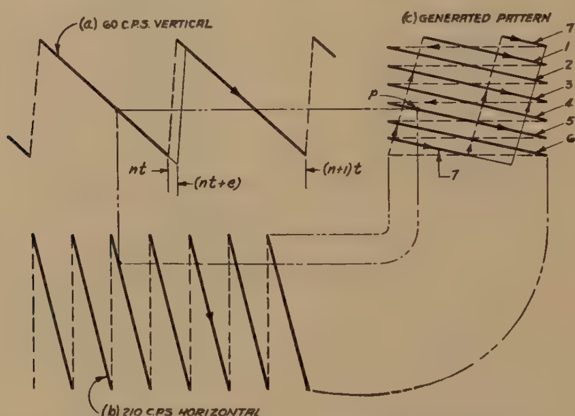


Fig. 5—Generation of interlaced scanning pattern by odd-line method.

plies, makes use of an odd number of horizontal scanning lines for each two vertical scanning cycles. For example, the condition now considered optimum for a video frequency band of 750 kilocycles is 243 lines in the complete picture, the frame frequency being thirty and the field frequency being sixty cycles. This makes the horizontal scanning frequency 7290 cycles and the lines per vertical deflection cycle $121\frac{1}{2}$. The half line left over at the end of each vertical cycle causes the alternate vertical deflection cycles to start 180 degrees apart with respect to the horizontal deflection cycle. Fig. 5 indicates how this will produce the interlaced effect. The lines 1, 3, 5, etc., are scanned during odd vertical saw-tooth cycles, whereas lines 2, 4, 6, etc. are scanned during the even vertical saw-tooth cycles. The arrows in the figure indicate the path of the scanning beam. Geometric construction for any point p in the pattern is shown by the broken line. (The light solid lines show a variaton in timing to be discussed later.)

Any one line is repeated only thirty times per second but no line flicker is perceptible because of the extremely small area occupied by a single line, and because of the small angle subtended at the eye by a single line. From data previously presented,[2] these factors are known to reduce flicker. Two or more alternate lines cannot co-operate to produce a thirty-cycle flicker by combining their area because if the eye includes more than one line, the intermediate lines will be unavoidably seen and the eye is subjected to the sixty-cycle alternating light effect as produced by the picture acting as a whole. One exception to this is possible but not probable and would occur when the subject matter of the transmitted picture chances to contain horizontal lines which agree

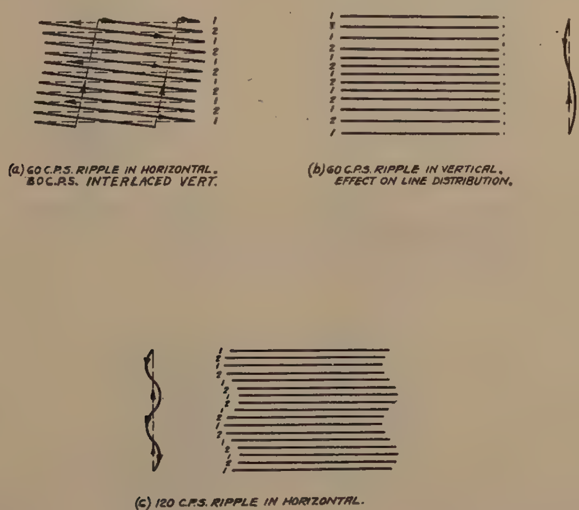


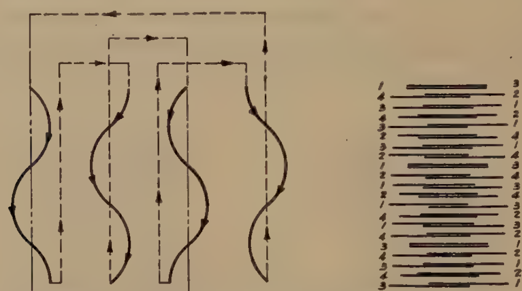
Fig. 6—Effect of ripple on thirty- to sixty-cycle interlaced pattern.

in position with the scanning lines such as to cause alternate lines to be dark over an appreciable area of the screen. Tests with this method of scanning have proved very satisfactory from the point of view of flicker for any ordinary picture subject matter.

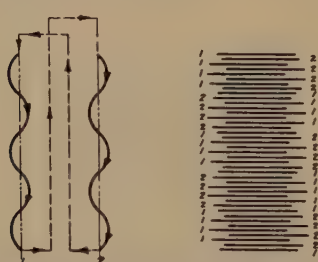
One slightly objectionable optical effect is noticeable in interlaced scanning pictures when objects in the scene move rapidly. If the motion is horizontal, the edges of the object appear to be jagged. This is due to the fact that a moving object is transmitted as a rapidly changing series of "stills" and that each alternate "still" is composed of only one set of alternate lines and that each "still" is slightly displaced horizontally with respect to the one preceding. On the other hand, the motion is actually portrayed more accurately by the thirty to sixty interlaced scanning than with thirty-frame progressive scanning, since the mov-

ing object is shown in sixty positions per second instead of thirty. This gain, however, is considered not to be of practical value.

When an object in the scene moves vertically, the apparent jagged edges of the object are not evidenced but the entire object may appear to be transmitted by a system having only half the total number of lines. This effect is complete and at its worst only if the motion is at the rate of one line pitch per one sixtieth of a second or an integer



(a) SCHEMATIC OF 60 C.P.S RIPPLE ON 24-48 C.P.S. INTERLACED PATTERN SHOWING RELATIVE PHASES FOR 4 VERTICAL DEFLECTION CYCLES. HORIZONTAL LINES SHOW EFFECT ON HORIZONTAL DEFLECTION. THE PAIRS OF LINES SHOWN CLOSELY ASSOCIATED ARE ACTUALLY SUPERIMPOSED.



(b) SCHEMATIC OF 120 C.P.S. RIPPLE ON HORIZONTAL FOR 24-48 C.P.S. INTERLACED PATTERN. FOR 1 COMPLETE CYCLE.

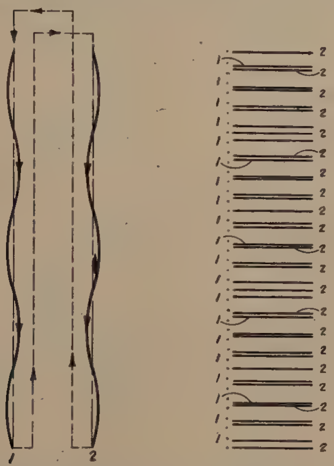
Fig. 7

multiple thereof. This loss of detail in a moving object is largely offset by the well-known fact that moving objects require less resolution in order to be understood, and that the eye cannot resolve minute detail in moving objects.

The effects of sixty-cycle and 120-cycle ripple upon interlaced patterns is shown in Figs. 6, 7, and 8. In the case of thirty- to sixty-cycle interlaced scanning (Fig. 6) the effects are very much the same as for thirty-cycle progressive scanning. The lines are displaced according to the sine law horizontally as shown in Figs. 6(a) and (c), and vertically as shown in Fig. 6(b). However, adjacent lines of the even and odd vertical deflections are all displaced similarly so that slight fixed

distortion of the picture is the only ill effect. Such distortion may have a magnitude equal to several times the line pitch for a 243-line picture without being serious.

For a twenty-four- to forty-eight-cycle interlaced pattern, Fig. 7(a) shows that four vertical deflection cycles are required for a complete recurrence of sixty-cycle ripple, and Fig. 7(b) shows that two are required for 120-cycle ripple. The horizontal lines in these figures are numbered to show to which deflection cycle they belong. (The return lines are omitted and the magnitude of the ripple effect is exaggerated to simplify illustration.) Fig. 8 shows the vertical displacement due to



SCHEMATIC OF 120 C.P.S. RIPPLE ON 24-48 C.P.S. INTERLACED PATTERN. EFFECT ON LINE DISTRIBUTION.

Fig. 8

120-cycle ripple. It is apparent that adjacent even and odd lines will be displaced in opposite directions, thereby causing serious loss of detail if the relative displacement has magnitudes comparable to the width of a picture element. (Such magnitudes of displacement would be unobservable in the case of the thirty to sixty interlaced pattern due to the similarity of displacements of adjacent lines.) The effect of ripple on the twenty-four- to forty-eight-cycle patterns is further objectionable in that horizontal displacement of lines causes the objects in the transmitted scene to appear to have jagged edges. The vertical displacement causes severe "pairing" of the lines in certain portions of the picture thereby destroying the benefits of interlacing and causing these portions of the picture to appear particularly coarse in structure by contrast with other portions. No figure shows the vertical displacement for sixty-cycle ripple but the net effect is very similar to that shown for 120-cycle ripple.

The odd-line method of producing interlacing as described requires only uniform saw-tooth wave shape deflection in which the vertical and horizontal scanning frequencies bear to one another a fixed ratio which is a whole number plus one half. To maintain these simple relations, however, presents several problems, mostly arising from the fact that the alternate vertical deflecting cycles start at different times with regard to the horizontal deflections. These starting times are shown as t_0 and $(t_0+1/60)$ seconds in Figs. 9(a) and (b). The two waves shown

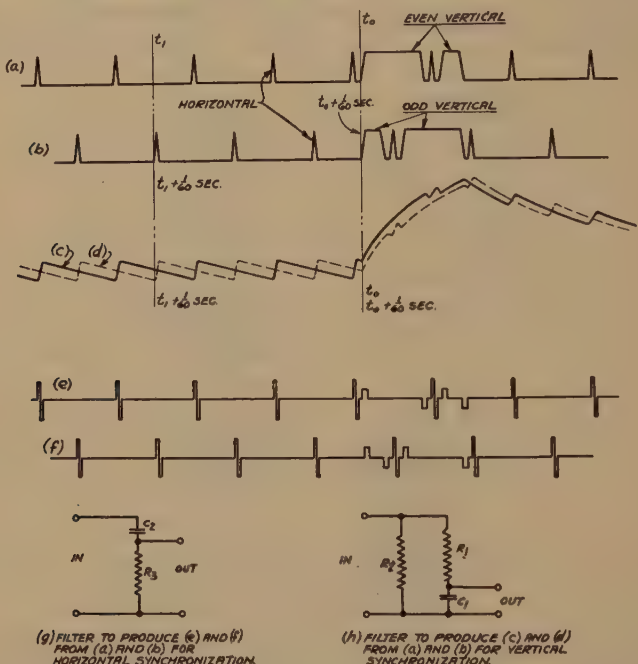


Fig. 9

are two different sections of the synchronizing wave, occurring alternately one sixtieth of a second apart. The "even vertical" synchronizing impulse occurs immediately following a horizontal synchronizing impulse while the "odd vertical" starts about midway between two "horizontals." Both vertical impulses have sections removed in order to accommodate "horizontals" that occur during the "vertical" in order that horizontal synchronization may be maintained continuously.

In one workable system, the synchronizing impulses Figs. 9(a) and (b) are transmitted by the picture transmitter and separated from the picture signal in the receiver by virtue of their greater amplitude of

transmission.[1] The impulses are then impressed upon a filter substantially as shown in Fig. 9(g) in order to obtain the impulses (e) and (f) for synchronization of the horizontal deflecting circuit. The function of the filter is primarily that of differentiation so that the output wave has magnitudes which depend upon the slopes of the input wave. The blocking oscillator for the horizontal deflecting circuit is not responsive to the small portion of the "vertical" impulse present in the wave (e) and (f) due to their timing as well as their reduced amplitude.

The impulses of Figs. 9(a) and (b) are also impressed upon the filter of Fig. 9(h) to obtain the alternate waves such as Figs. 9(c) and (d) for the synchronization of the vertical deflecting circuit. This filter primarily integrates the impulses of the input wave as the output is

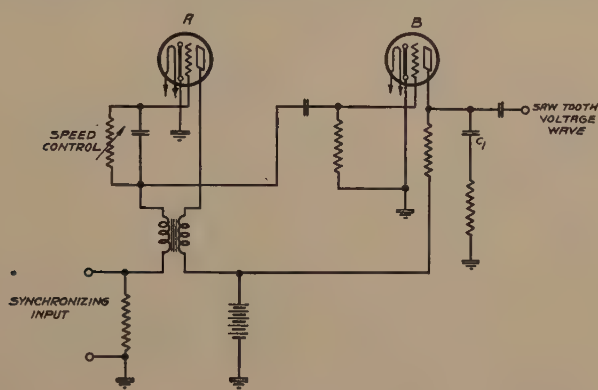


Fig. 10—Blocking oscillator and saw-tooth generator.

the voltage across a condenser C_1 charged through a resistor R_1. The "horizontal" impulses have amplitudes greatly reduced in comparison with the "vertical" impulses due to their shorter duration. However, a study of (c) and (d) shows that the "vertical" impulses differ considerably from one another in their amplitude corresponding to any short interval after t_0 or $(t_0+1/60)$ seconds on the increasing side of the peak. This difference, which is due to the dissymmetry of the "horizontals" with respect to the two "verticals," tends to cause the oscillator used in the vertical saw-tooth deflecting circuit and synchronized by (c) and (d), to operate at a nonuniform speed. It also tends to make the alternate even and odd saw-tooth strokes differ slightly in amplitude as will be evident from a study of the method of generation of the saw-tooth wave by the circuit of Fig. 10. The tube A and associated circuit comprise the blocking oscillator which generates impulses such as the wave of Fig. 11(g), which has an amplitude about fifty times as great as the synchronizing wave, Figs. 9(c) and (d). In

the circuit it is apparent that the synchronizing wave, though intentionally impressed upon only the oscillator, is also impressed through the transformer winding upon the grid of the discharge tube B, along with the output of the oscillator. The amount of each discharge of the condenser C_1 is dependent upon the amplitude, shape, and duration of the impulse supplied to the grid of tube B during the discharge time. The difference in waves (c) and (d) of Fig. 9 causes these factors to differ slightly for the even and odd impulses, thereby causing the alternate discharges of condenser C_1 to differ slightly in amplitude.

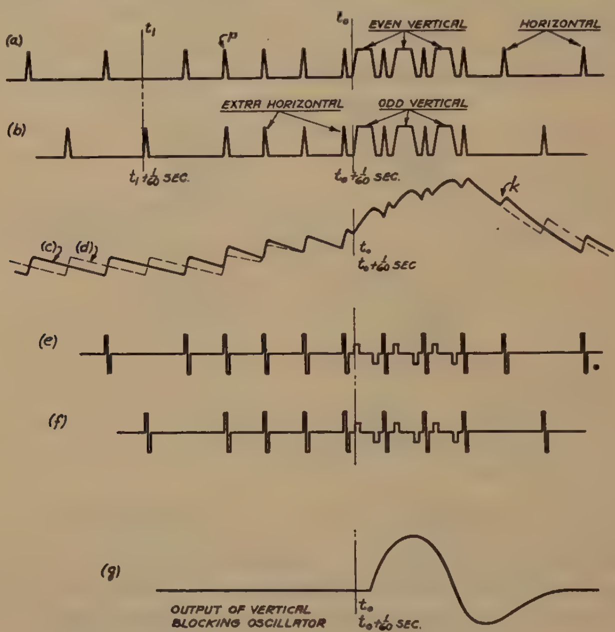


Fig. 11—Synchronizing waves for odd-line interlaced scanning.

The net effect of this is a slight vertical displacement of the horizontal lines of the odd vertical cycles with respect to those of the even cycles. (Further explanation of this phenomenon will be given below in connection with "even-line" interlacing.) The displacement will be so great as to destroy completely the benefits and appearance of interlacing if the difference in magnitude of alternate discharges differs by as much as 0.41 per cent. A lesser amount will cause the lines to be grouped in pairs.

Extremely accurate timing of the alternate vertical saw-tooth generating discharges is not directly necessary for sufficiently uniform spacing of the lines in interlaced scanning. In Fig. 5 the light solid

lines show the change caused in the generated pattern when the even discharges begin at a time $(nt+e)$ instead of time nt as would be required for perfect timing. As drawn, the interlacing is not impaired, the assumption being made that all discharges are identical in magnitude. Actually the delayed discharge would be slightly greater due to a slightly greater average voltage on the plate of the discharge tube during the discharge. (For a screen-grid type discharge tube, this discrepancy would be less.) However, indirectly nonuniform timing has a large effect upon the spacing of the lines since it alters the time relation of the synchronizing impulse to the discharge time and thereby changes the effectiveness of the synchronizing impulses as a contributor to the magnitude of the discharges.

Several methods have been developed which satisfactorily overcome the dissymmetry of the "integrated" synchronizing impulses. One method is to make the synchronizing impulses identical in the region of the even and odd vertical synchronizing impulses. This is accomplished by the arbitrary introduction at the transmitter of additional impulses similar to the horizontal synchronizing impulses. They are located midway between each two regular horizontal synchronizing impulses for an interval of a few line periods before and during the vertical synchronizing impulses, as shown in Figs. 11(a) and (b). The vertical synchronizing impulses are interrupted at half-line period intervals in order to accommodate the regular and the additional horizontal synchronizing impulses. The first additional impulse "p" may be of different duration than the others in order to compensate partially the charge in the "integrating" condenser for the necessary dissymmetry preceding the region of additional impulses. The partially integrated waves of the even and odd vertical synchronizing waves are respectively shown at (c) and (d), Fig. 11. They are practically identical during the occurrence of the oscillator impulse (g).

Of course, the "extra horizontal" impulses pass through the horizontal impulse selecting filter the same as do the regular "horizontal" impulses, Figs. 11(e) and (f). However, due to their timing, the blocking oscillator of the horizontal deflecting circuit does not respond to them. Use of the modified synchronizing wave has resulted in better operation with less critical adjustment of the vertical deflection speed control and of the amount of impressed synchronizing signal.

Another method for overcoming the effect of dissymmetry of the synchronizing waves upon interlacing involves the use in the receiver of an additional vertical frequency blocking oscillator, which serves as a "buffer." This oscillator synchronizes on the "integrated" synchronizing wave and in turn supplies synchronizing impulses having a

higher degree of uniformity in amplitude to the blocking oscillator used to produce the saw tooth of voltage. However, the alteration of the synchronizing wave is the preferred solution, since it entails no additional receiver equipment.

Dissimilar vertical saw-tooth waves may also be caused by "cross talk" between the horizontal and vertical deflecting circuits in the receiver. This may be overcome satisfactorily by moderate shielding and the exercise of care in the location of parts.

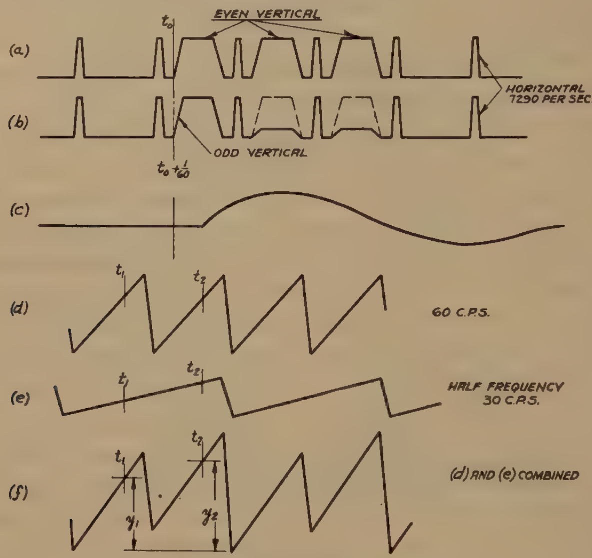


Fig. 12—Even-line interlacing.

Even-Line Interlaced Scanning

The even-line method of interlacing involves the utilization of the phenomenon which was just described as the greatest handicap of proper interlacing by the odd-line method, namely, dissimilar even and odd vertical saw-tooth generating discharges. For the sake of analysis a train of discharge impulses of sixty-cycle recurrence frequency but having alternate impulses of unequal amplitudes, may be considered to consist of two separate trains of discharge impulses as follows, acting simultaneously: One uniform train of sixty-cycle impulses and one uniform train of thirty-cycle impulses having an amplitude equal to the difference in the alternate discharges of the original train of impulses. These two trains respectively produce the saw-tooth waves of Figs. 12 (d) and (e) which combined make the wave of Fig. 12(f).

For the purpose of illustration, the wave of Fig. 13(b) represents a horizontal deflecting saw-tooth wave of four times the frequency of

the vertical wave. Fig. 13(c) represents the scanning pattern traced on the iconoscope or kinescope when deflected simultaneously by the vertical and horizontal waves of Figs. 12 (a) and (b). The arrows on the lines show the direction of travel of the scanning beam. This pattern readily indicates that an interlaced effect will be obtained because the lines for adjacent vertical scannings are alternated in position. Also the fact that the similar lines for adjacent vertical scannings will not fall upon one another will be clear from Fig. 12(f) since y_1 and y_2 corresponding to t_1 and t_2, respectively, are not equal when t_1, t_2, etc., represent the period of the wave of Fig. 12(d). If the amplitude of the

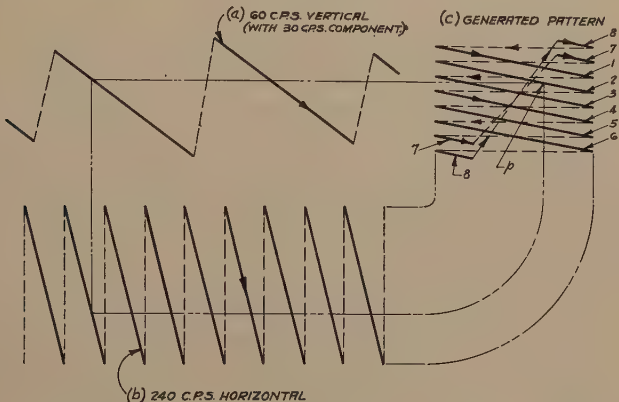


Fig. 13—Generation of interlaced scanning pattern by even-line method.

wave of Fig. 12(e) is altered, $(y_2 - y_1)$ will be changed and the spacing between the "even" and "odd" lines will be altered. This leads to the major defect of this method of interlacing, namely, that the value of the difference between the amplitude of the alternate discharges is somewhat critical for perfect interlacing, which might necessitate the use of an additional control on the receiver.

Figs. 12(a) and (b) show the even and odd sections of the synchronizing wave used. The actual synchronization of the vertical oscillator is maintained by the first-occurring portion of the wave, which is identical for the even and odd cycles. However, the impulse Fig. 12(c) generated by the oscillator occurs slightly later and includes part of the time in which the even and odd portions differ as shown by the dotted line Fig. 12(b). This synchronizing wave is used for two purposes, as follows: It is filtered to remove the horizontal impulses substantially, as was described for the odd-line method of interlacing, and then impressed upon the vertical oscillator for synchronizing. The synchronizing wave (unfiltered) is also impressed upon the cathode circuit of the discharge tube, while the output of the oscillator is impressed upon

the grid to cause the discharge tube to pass plate current intermittently and thereby generate the vertical saw-tooth wave. The plate conductivity of the discharge tube is a function of the cathode potential as well as the grid potential. Therefore, since the oscillator output wave Fig. 12(c) occurs partly during the interval in which wave portions of Figs. 12(a) and (b) are different, the alternate discharges will be different. The difference in area under the "even" and "odd" portions of the wave, which causes the inequality of the alternate discharges, needs to be only a fraction of one per cent of the area under the oscillator impulse as applied to the grid circuit of the discharge tube.

Other Problems Associated with Interlaced Scanning

For a picture containing such a great number of lines, difficulty was experienced in the mechanical accuracy of construction of a synchronizing signal generator of the rotary type. An electrical synchronizing signal generator has been developed, which consists of special multivibrator and wave-shaping circuits synchronized with the sixty-cycle power supply. Further description of this signal generator is beyond the scope of this paper.

In order to use standard thirty-five-millimeter film for television subject material it was necessary to design the film projector in such a way that with the film running at twenty-four frames per second, it could be scanned at a rate of thirty complete frames per second or sixty exposures of the iconoscope plate per second as required for interlacing. By using a specially designed cam-driven oscillating mirror type projector having a very quick return, it is possible to maintain the image of the even frames stationary for substantially two sixtieths of a second, or for two vertical scannings, and the odd frames stationary for substantially three sixtieths of a second, or for three scannings. By this procedure two frames of film are projected in five sixtieths or one twelfth of a second, which gives the required twenty-four frames per second.

Complex Scanning Patterns

The use of more complex types of interlaced scanning has been suggested. One of these, known as "triple interlacing" has been used considerably in mechanical scanning in which a three-spiral disk was used. Triple interlacing has also been used in a laboratory installation of a cathode-ray system. In the latter case the interlacing was accomplished by a method using the same principle as the "odd-line" method described, the difference being that the horizontal scanning frequency was made a whole number plus one third (instead of one half) times the vertical saw-tooth frequency. In triple interlaced scan-

ning, the screen may be considered to be divided into groups of lines, each group consisting of three adjacent lines numbered 1, 2, and 3. Then the scanning cycle would be, first, all lines numbered 1; second, all lines numbered 2; and third, all lines numbered 3. If triple interlacing is to be useful, the most promising vertical scanning frequency for cathode-ray use is sixty, which gives a picture repetition rate of twenty. At this low frequency the individual lines exhibit appreciable flicker and also the screen produces a disagreeable effect of moving or "crawling." This effect is due to the fact that if the eye moves over the screen vertically downward at a certain constant speed, it will rest on line 1 of group one, while it is being scanned, then on line 2, then line 3, then line 1 of the next group, etc., while each one is being scanned. Actually, if the attention of the eye is permitted to follow this progression, the screen will appear to become very coarse and to have only one third of the total number of lines. This effect, which renders triple interlacing very objectionable, is practically absent on the "double" interlaced screen, it being possible to observe the effect only by special effort of the observer. The marked difference between the double and triple interlaced pattern is due to the higher repetition rate of the individual lines and the larger ratio of the width of a line to the width of the group. (There are two lines in each group for double interlacing.) For quadruple or higher interlacing, it is believed that these ill effects would become more serious.

Conclusion

In conclusion, the writers believe that double interlaced scanning with a frame frequency of thirty per second is the optimum known condition at the present time for alternating-current power supply sources of sixty cycles per second.

Since the minimum picture repetition rate for negligible flicker has been set at forty-eight pictures per second, it is interesting to compare the picture detail provided by progressive scanning at forty-eight pictures per second to that provided by interlaced scanning at thirty pictures per second, with a maximum video frequency of 750 kilocycles.

From the formula for equal horizontal and vertical detail,

$$a = \sqrt{2f/nRK}$$

where,

a = the number of scanning lines
f = maximum video frequency in cycles
R = aspect ratio $(=4/3)$
n = frame repetition rate
$K = 0.64$

the number of lines, a, is 192 for progressive scanning and 243 for interlaced scanning.

The effective number of picture elements, e, may be obtained by the formula, $e = a^2RK^2$, in which K^2 is a correcting factor required on account of the loss due to random details of the picture not coinciding with the scanning line as discussed by the writers in the paper previously cited. (The losses of picture elements due to scanning beam return time and synchronizing are neglected in this case for simplicity.) The effective number of picture elements is 20,200 for progressive scanning and 32,100 for interlaced scanning.

From the above it is seen that the progressive scanning provides only sixty-two per cent of the detail provided by "interlaced" scanning. One case in which thirty to sixty interlacing would not be optimum would be for receivers located in alternating-current power districts other than sixty cycles, receiving programs of a transmitter in a sixty-cycle area. In this case, the benefits of interlacing would be largely lost due to alternating-current ripple. (Extra precautions taken in shielding and filtering the power supply of receivers for such special conditions will reduce the ripple effects and might make the interlacing acceptable.) The example just given for transmitter and receiver in differing power supply frequency areas would represent only a negligible portion of the probable installations except when considering relayed programs. Where transmitter and receivers are located in the same power supply districts as regards frequency, interlacing can always be satisfactorily obtained by proper choice of operating characteristics; i.e., for fifty-cycle power source, a frame frequency of twenty-five per second and a field frequency of fifty per second. (For a fifty-cycle source, operation at the standard movie speed of twenty-four frames per second would be satisfactory.)

In direct-current power districts it is at present necessary to use a converter with the alternating-current type of receiver. With the frequency of the converter adjusted within one or two cycles of the power supply at the transmitter, all of the advantages of interlacing are had. Interlacing also raises the minimum video frequency requirement in the ratio of two to one, since it increases the field frequency, thereby reducing the amplifier difficulties and cost at both transmitter and receiver.

Acknowledgment

The authors express appreciation to Mr. W. A. Tolson for assistance in the work pertaining to the relation of the frame frequency and the alternating-current power supply, and to Messrs. W. J. Poch and J. P. Smith for assistance in producing and testing the effect of various synchronizing wave shapes.

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A PROPOSED WATTMETER USING MULTIELECTRODE TUBES*

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FOR direct measurement of small amounts of power, or for measurement of power over a wide range of frequencies, dynamometer wattmeters are unsatisfactory, and vacuum tube wattmeters are often resorted to.

Hitherto vacuum tube wattmeters have depended for their operation on the use of tubes having a square-law characteristic, connected in a push-pull circuit so as to avoid indications caused by current or voltage alone. Such a device was described by H. M. Turner and F. T. MacNamara in 1930[1] and was patented by E. Peterson in 1926.[2] References to this type of wattmeter may also be found in standard works on radio-frequency measurement.[3,4]

An attempt to use a multielectrode tube with coplanar grids for the measurement of power was made by T. B. Wagner.[5] While the device developed was satisfactory as a voltmeter or ammeter, it must be discounted as a wattmeter, for it is clearly brought out in Wagner's article that readings may be obtained by applying voltage or current alone, a condition of zero power. Moreover, it is stated that the output of the tube used is proportional to the sum of the effects of voltage and current if applied separately.

The author wishes to propose a new method of measuring power through the use of multielectrode tubes.

It may be shown of some multielectrode tubes that under certain conditions the application of alternating voltage to two controlling elements or grids results in a change in the direct-current component of the plate current proportional to the product of the voltages at the grids and the cosine of their phase angle. Hence such tubes may be used in the construction of a vacuum tube wattmeter.

* Decimal classification: R240. Original manuscript received by the Institute, September 30, 1935; revised manuscript received by the Institute, January 2, 1936.

[1] H. M. Turner and F. T. MacNamara, PROC. I.R.E., vol. 18, pp. 1743–1747; October, (1930).
[2] Patent No. 1,586,553, E. Peterson, Bell Telephone Laboratories, June, 1926.
[3] Terman, "Measurements in Radio Engineering," pp. 31–32.
[4] Hund, "High-Frequency Measurements," pp. 302–303.
[5] T. B. Wagner, *Elec. Eng.*, vol. 53, pp. 1621–1623; December, (1934).

Of the tubes investigated by the author, the mixer type of tube, such as the type 2A7 or 6A7, seemed best adapted for use as a wattmeter tube. The disposition of electrodes in a type 2A7 or 6A7 tube is shown in Fig. 1. G_2 and G_3 are held positive at a fixed potential above that of the cathode, and G_1 and G_4 are biased to be negative with respect to the cathode. Alternating-current potentials proportional to voltage and current are then applied to G_1 and G_4.

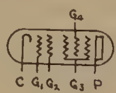

Fig. 1

In such a case, the current passing through G_1 is dependent only on the potential of G_1, being independent of the potential of G_4. The

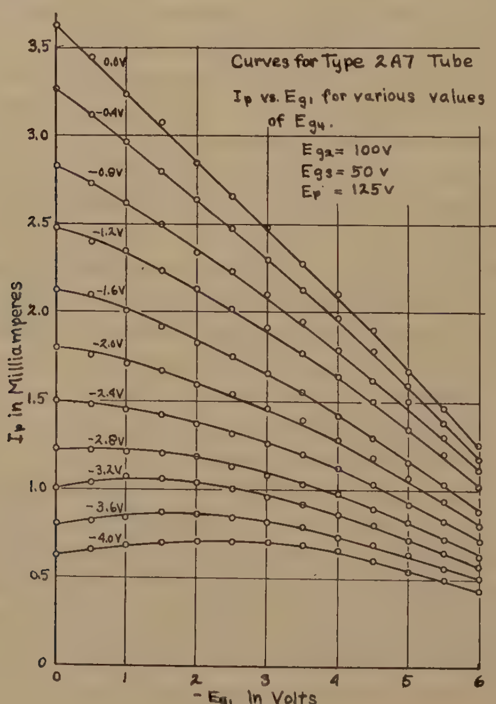

Fig. 2

proportion of this current reaching the plate is, however, dependent only on the potential of G_4. Thus if the characteristics are linear for both G_1 and G_4, we may expect that the plate current will be represented in the form

$$I_p = A\,E_{G1} + B\,E_{G1}E_{G4} + CE_{G4} + D \qquad (1)$$

where A, B, C, and D are constants.

The first and third terms are necessarily of a purely alternating-current nature and would not register on a direct-current meter in the plate circuit. The direct-current component of the product term is proportional to the power, since E_{G1} and E_{G4} are proportional to alternating voltage and current.

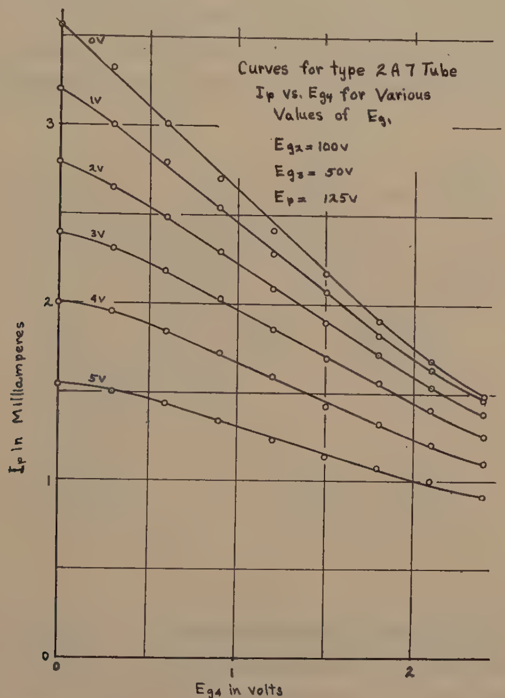

Fig. 3

The actual static characteristics, experimentally obtained by the author, for a 2A7 tube are shown in Figs. 2 and 3. Fig. 2 is a plot of I_p vs. E_{G1} for various values of E_{G4}, and Fig. 3 is a plot of I_p vs. E_{G4} for various values of E_{G1}.

If a region can be located such that within it I_p vs. E_{G1} for various values of E_{G4} can be represented by a family of straight lines which, extended, pass through a common point, and if in this region I_p vs. E_{G4} is, for some value of E_{G1}, a straight line, then the plate current of the tube can be represented by (1) and the tube may be used as a wattmeter tube.

Figs. 2 and 3 should, if sufficiently accurate, enable us to find such a region if it existed. Unfortunately, there is no large region of linear variation of I_p with E_{G1}, although something approximating such a region may be found. Further, curves such as these can hardly be accurate enough to give final evidence of the degree of linearity, and can best be used to show under what operating conditions linearity may be expected.

Variation of the tube characteristic from linearity is best detected by applying various alternating-current potentials to one grid alone and noting the variation in the direct-current component of the plate current due to "rectification." Making such tests, the author found that

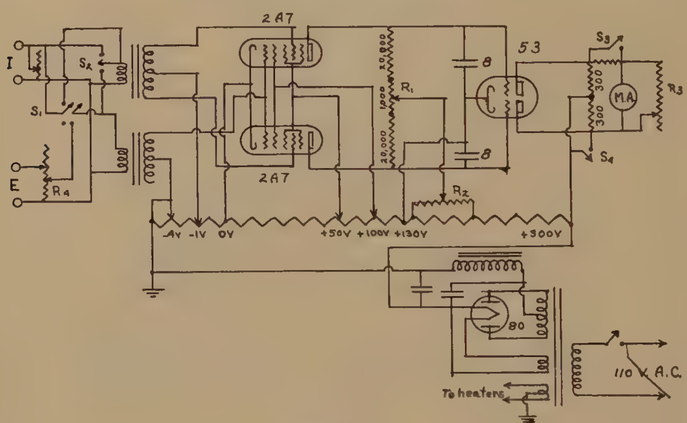

Fig. 4

the variation in the direct-current component of the plate current caused by an alternating voltage applied to G_1 alone was great enough to preclude the use of a single tube in a wattmeter, since "rectification" would necessarily result in readings due to current or voltage alone.

Linearity is much more closely approached in the variation of I_p with E_{G4} alone. The best result obtained was a change in the direct-current component of I_p caused by the application of an alternating voltage to G_4 alone not more than three per cent of that for the same alternating voltage applied to both G_1 and G_4.

It is possible that tubes similar to the type 2A7 but having more nearly linear characteristics could be developed. This presents an interesting field of investigation which was closed to the author because of lack of equipment.

In view of the nonlinearity of the characteristics of the tubes available, a push-pull type of circuit was adopted to check experimentally

the operation of a wattmeter constructed on the principles outlined above. The circuit is shown in Fig. 4. The apparatus was built in a self-contained form with variable shunts and voltage dividers to cover a range from 0.01 to 1.00 amperes and 1 to 200 volts, and with switches to allow for the use of the instrument as a square-law voltmeter or ammeter as well as a wattmeter. A direct-current amplifier was provided to allow the use of a rugged one-milliampere meter in place of the microammeter which would otherwise have been required.

The instrument was tested at fifty cycles. The curves shown in Figs. 5 and 6 demonstrate that at this frequency the instrument behaves as a wattmeter should. Fig. 5 is a plot of deflection as a watt-

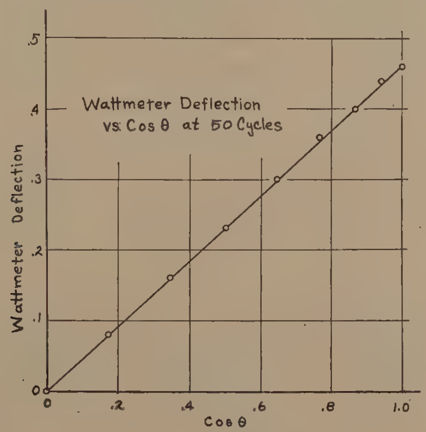

Fig. 5

meter with constant voltage and current vs. cosine of the phase angle, and shows that the instrument behaves correctly as far as power factor is concerned. In obtaining this experimental plot the author used an iron-core phase shifting transformer, such as is used in the calibration of watt-hour meters. Fig. 6 is a plot of indicated amperes (the square root of scale reading) for the instrument as an ammeter vs. actual amperes as measured by an ammeter. This illustrates that the deflection is actually proportional to the product of the voltages at the grids, in this case each proportional to the current.

With this circuit, the frequency limitation is of course that of the transformers. Obviously, the phase shift of the transformers does not matter as long as the transformers in the voltage and current sides have equal phase shifts, as they will if they are identical. It is important, however, that the ratio of transformation does not vary with frequency and that the transformers do not in any way distort the wave

form. In the present state of the art there should be small difficulty in obtaining transformers having a substantially constant ratio of transformation over the audio range, and giving no appreciable distortion of wave form. With such transformers the device would function as satisfactorily over the entire audio range as it does at fifty cycles.

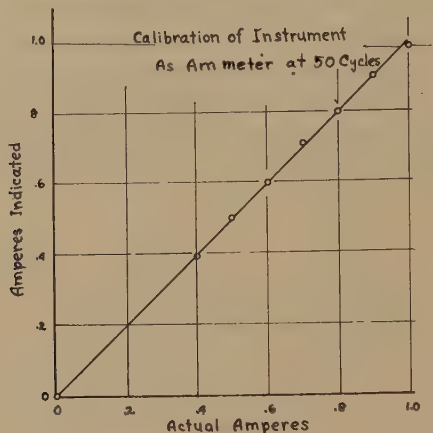

Fig. 6

It should be noted that such an instrument in conjunction with an audio-frequency oscillator would serve admirably as an harmonic analyzer of a zero-beat type, such as described by C. G. Suits.[6]

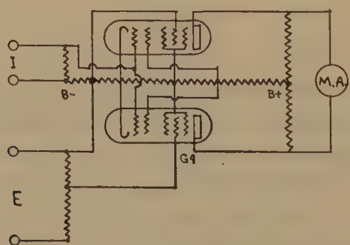

Fig. 7

In extending the range of the wattmeter to cover radio frequencies, transformers must, of course, be avoided. As has been explained, a single tube could be used were its characteristics for both control grids sufficiently linear. Provided the characteristic of one grid only is substantially linear, as in the case of the type 2A7 tube, the circuit suggested in Fig. 7 should provide a means of reasonably accurate power

[6] C. G. Suits, Proc. I.R.E., vol. 18, pp. 178–192; January, (1930).

measurement without the use of transformers. This circuit is symmetrical with respect to the applied current; hence deflections due to current alone are avoided. It is not symmetrical with respect to the applied voltage, but sufficient linearity of the characteristic for G_4 to avoid immoderate deflections caused by voltage alone is presumed. A certain limitation of this circuit, that one side of the current input and one side of the voltage input must be at the same radio-frequency potential, must be common to all vacuum tube wattmeters not making use of transformers. Direct-current isolation may be obtained through the use of condensers.

In conclusion, it may be asserted that present multielectrode tubes of the mixer type can be utilized to advantage as wattmeter tubes, and that with similar tubes having more nearly linear characteristics, the art of power measurement at low levels and varying frequencies might be considerably advanced.

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RADIO PANEL LAMPS AND THEIR CHARACTERISTICS*

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Introduction

THE original conception of a radio panel lamp or, in more popular terms a dial light, envisaged no more difficult problem than constructing a small light source of low power for illuminating the station markings on the opaque dials then used. The lamp was simply mounted in a position to permit the light to fall directly upon the surface of the dial.

The introduction of translucent dials presented no new problems except with the later forms of airplane dials which required more attention to be paid to the filament form and to the selection of bulbs so as to prevent streaky illumination of the dial.

In more recent years, however, the simple requirements of dial illumination have been displaced by the more stringent optical requirements of tuning meters of the shadow-producing type. Since it was desirable to retain the well-established physical form of the lamps as used for dial illumination, while at the same time designing them to meet the requirements of suitable tuning meter lamps, the deceiving simplicity of these lamps, more often than not, caused the radio manufacturers, as well as the lamp manufacturers no small amount of trouble.

It is the practice in such shadow meters to place the lamp directly behind a small aperture of predetermined size so that the light shines directly past the edges of a movable vane to cast a shadow of the vane on a translucent screen. The vane, which is actuated by a small magnetic coil, through which the tube current flows, will show a minimum deflection for the condition of resonance in the circuit and the shadow of the vane on the screen will, correspondingly, have a minimum width.

To have maximum contrast it is necessary for the edges of the vane to be sharply delineated on the screen and free from penumbra effects. This requires a special straight line form of filament known as the C-6. Precautions must be taken to place the filament in the bulb so that images reflected from the bulb walls will not pass through the aperture and act as secondary sources of light.

* Decimal classification: R380. Original manuscript received by the Institute, July 25, 1935. Presented before Tenth Annual Convention, Detroit, Mich., July 1, 1935.

Elements of Lamp Design

Tungsten Wire Properties—With drawn tungsten wire, the internal structure of the metal, after drawing, consists of very long crystals of tungsten. As the lamp is burned, these crystals break up

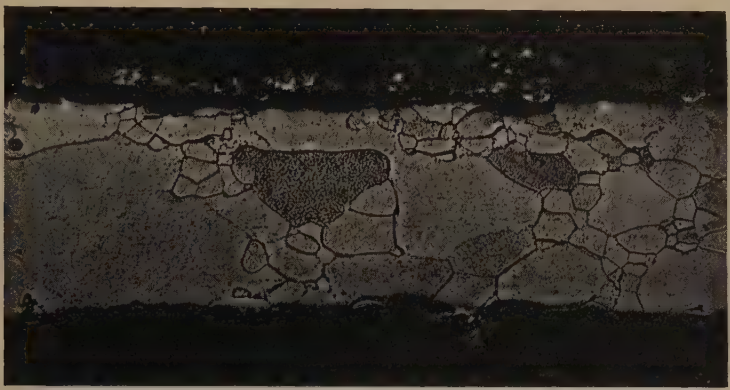

Fig. 1—Appearance of untreated tungsten wire after burning, showing brittle condition due to irregular crystal formation.

into smaller sections so that the filament is made up of millions of tiny blocks of tungsten. (Fig. 1.)

The advent of nonsag wire, in which very small amounts of either potassium or sodium are used, represented a very distinct improve-

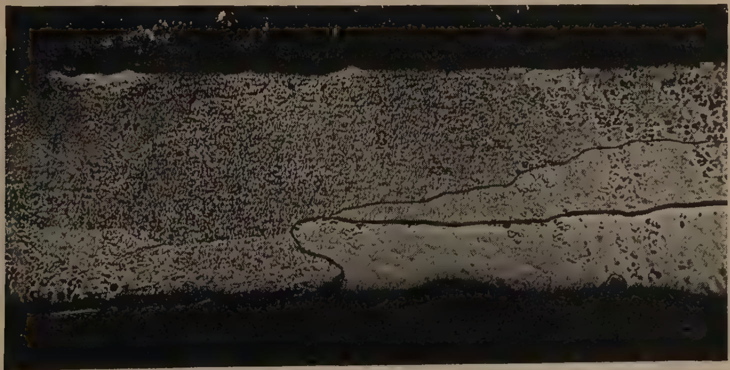

Fig. 2—Appearance of nonsag tungsten wire after burning, showing formation of long interlocking crystals which prevents wire from sagging.

ment in filament performance. In such nonsag wire, the crystals, which form as the filament is burned, interlock with each other like a jigsaw puzzle in such a manner as to improve the strength of the filament. (Fig. 2.) Such wire will withstand shock, although it will not withstand

severe vibration. This wire is used in about ninety-five per cent of all types of lamps made today and it is also used in the manufacture of radio panel lamps. While there is loud speaker vibration present in radio receivers, no difficulty is encountered on this score provided the lamp is not mounted on the speaker frame or where the speaker vibration can be transmitted directly to the lamp.

Another form of tungsten wire with altogether different characteristics is obtained by adding very small amounts of thoria to the tungsten. As the thoriated filament is burned, very small round-edged crystals are formed and because of the large number of such crystals throughout the cross section of any given size of wire, slippage of the crystals does not affect the filament in general. (Fig. 3.) Such wire is

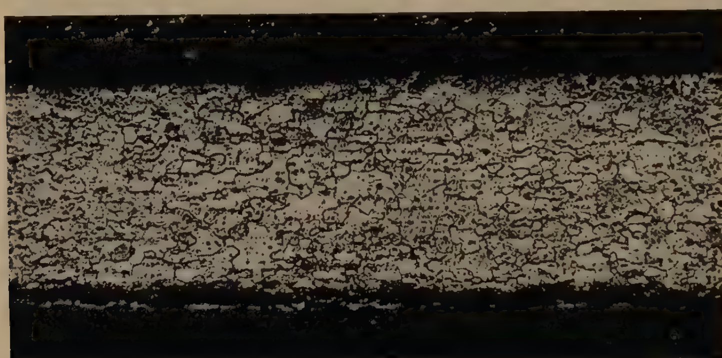

Fig. 3—Appearance of treated sag tungsten wire after burning, showing formation of minute irregular crystals.

known as sag wire and while it will withstand vibration very well, it is not so good under shock. It is the practice to use such wire for street railway lamps. Occasionally, when radio manufacturers thoughtlessly mount their sockets in such positions that they are subjected to the full effect of the speaker vibrations, it is necessary to come to their rescue by temporarily supplying lamps having sag-wire filaments, but with the result that other forms of trouble are encountered.

Filament Forms—There are several outstanding filament forms used in the design of radio panel lamps. The straight horizontal coil mentioned in connection with the lamps used for tuning meters is known as the C-6 form. It is somewhat difficult to obtain a perfect C-6 form in such low-priced lamps, where automatic coiling and mounting machinery must necessarily be used for high production, low cost work. In spite of such difficulty, however, very satisfactory results, which fully meet the requirements, are obtained with present methods.

The first form of coiled filament used in the early panel lamps was the arched, or C-2, type where the filament is mounted bow-shaped between the two copper lead wires. This filament has the advantage of casting light around the lead wires so as to prevent sharp shadows of the leads being cast on the illuminated dial. In some instances, it is necessary to regulate the degree of arching to obtain satisfactory results in practice. This filament form is still used in certain types of lamps, but it is only suitable for dial illumination, and since the trend is very definitely toward a single filament design, which would be satisfactory for all radio purposes, it may be superseded by the C-6 form so necessary in shadow meters, and which can also be applied to dial illumination by positioning the lamp so that the filament is parallel to the dial.

In those special forms of lamps used in battery-operated receivers, where the current is of the order of sixty milliamperes and the voltage is also low, a straight wire filament form, known as the S-2, is used because the shortness of the filament almost precludes coiling.

Lamp Bulbs—Probably more for the reason of having a distinguishing bulb type than for any other reason, the standard form of bulb for radio panel lamp service is the small tubular type known as the T-3$\frac{1}{4}$, measuring approximately three and one-quarter eighths of an inch in diameter. Consideration has been given at various times to other bulb forms, notably the globular type and a special adaptation of the tubular type having sloping side walls after the fashion of an acorn. Here again, standardization, which means low cost and low price, played an important part so that all panel lamps for use in home receivers are of the T-3$\frac{1}{4}$ bulb form. In the automotive field, however, where the globular or G type bulb has been in use for many years and also because of its shorter over-all length to facilitate mounting in the restricted steering column controls, this bulb is now standard in lamps used in most receivers.

Lamp Bases—Until recently, the only lamp base used for radio panel lamps was the miniature screw base. As receiver design improved, however, and also due to the use of the tuning meter, it became necessary to consider another form of base in order to remove many of the difficulties inherent in the screw-base type. Approximately one-third of lamp outages in receivers is due to the lamps vibrating loose in the sockets, thereby setting up growling noises. Various manufacturers in an effort to correct this condition, found in the screw-base lamps, were using such means as crimping the socket and then forcing the lamp into the socket with pliers so as to lock the

lamp in position. This not only made it extremely difficult to replace burned-out lamps but also resulted in cracking from twenty to twenty-five per cent of the glass bulbs below the top line of the base where such cracks could not easily be detected. Those lamps then failed early in life due to leakage trouble.

To remedy this situation, a miniature bayonet base was made available recently which is just like that employed on the common automobile types of lamps, except that it is smaller. The use of this miniature bayonet base is expected to remove many of the troubles formerly encountered with the old types of screw base and its use is therefore urged upon receiver manufacturers.

Lamp Inspections—It has been found necessary, in manufacturing these lamps, to use special means to insure obtaining a good clamp between the ends of the filament and the copper lead wires. Soft copper is used to permit the filament ends to be buried in the copper without fracturing the tungsten wire. A hot clamp is used to obtain the effect of a weld. Each individual lamp is then tested in an amplifying outfit whose level of amplification is fixed to make sure that those lamps having open circuits will be weeded out. In addition to the many other regular inspections conducted as a routine matter, these special precautions are required so as to prevent the lamp from creating extraneous noises in the receiver. It is also necessary to obtain lamp bulbs which are free from seeds, chords, and mold marks, which imperfections produce shadows.

Lamp Characteristics

In any industry which involves the manufacture of hundreds of millions of units, standardization necessarily assumes a very important rôle so that lamp manufacturers look upon standard types as being the only ones which really need be used. Only those lamps are placed in the standard schedules whose quality is definitely assured through endless tests and whose distribution is widespread. Such lamps, naturally, carry the lowest price. Nevertheless, many other types of lamps are made available to meet all requirements.

The types of radio panel lamps now available are Mazda lamps Nos. 41, 43, 44, and 46 for home receivers and Mazda lamps Nos. 50, 51, and 55 for automotive receivers. (Table I.) In addition, there are a number of very special types whose use is restricted to special applications of limited distribution.

Wherever possible, standard types of lamps should be used as their performance is well known by reason by numerous tests and because their quality is assured by manufacturing experience acquired through continuous production.

TABLE I
POPULAR TYPES OF RADIO PANEL LAMPS

Mazda Lamp No.	Volts	Amperes	Bulb	Base	Life	Service
40	6.3	0.15	T-3¼, clear	Min. Screw	3000 hrs. at 6.3 v.	Radio dials
41	2.5	0.50	T-3¼, clear (also frosted)	Min. Screw	3000 hrs. at 2.5 v.	Radio dials
43	2.5	0.65	T-3¼, clear	Min. Bayonet	3000 hrs. at 2.5 v.	Radio dials and tuning meters
44	6.3	0.25	T-3¼, clear	Min. Bayonet	3000 hrs. at 6.3 v.	Radio dials and tuning meters
46	6.3	0.25	T-3¼, clear	Min. Screw	3000 hrs. at 6.3 v.	Radio dials and tuning meters
50	6–8	C.P. 1.0 Amp. 0.20	G-3½, clear, red or green	Min. Screw	1000 hrs. at 7.5 v.	Radio dials
51	6–8	C.P. 1.0 Amp. 0.20	G-3½, clear, red or green	Min. Bayonet	1000 hrs. at 7.5 v.	Radio dials
55	6–8	C.P. 1¾ Amp. 0.40	G-4½, clear	Min. Bayonet	500 hrs. at 6.5 v.	Radio dials

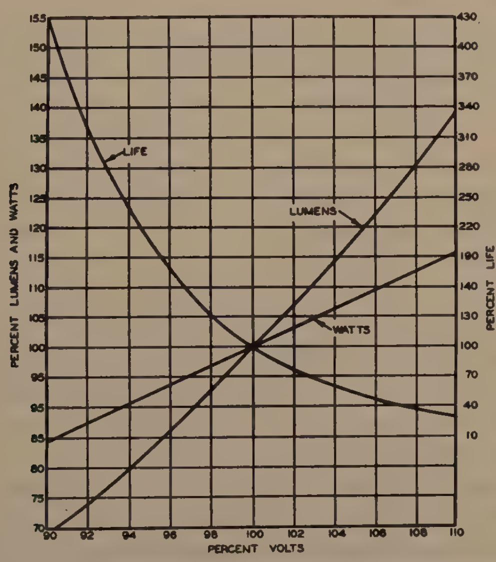

Fig. 4—Variation of life, lumens, and watts with applied volts.

The life of these lamps will necessarily vary with the applied voltage and it can be said that a change of one per cent in applied volts will result in a change of approximately ten per cent in life. (Fig. 4.)

While all lamps are designed for an average life rating, it does not necessarily follow that each lamp will meet the given figure. On the contrary, in any given group of lamps, a varying rate of lamp failure will be found and this rate of failure is reasonably definite for large groups of any given type of lamp. The average life of any appreciable group of lamps will meet the designed figure due to the fact that some lamps burn considerably beyond the designed life figure and com-

pensate for those which fail before the designed life figure has been reached. For example, in radio panel lamps, about thirteen per cent of the lamps will have failed at fifty per cent of designed life and about fifty-four per cent of the lamps will have failed at one hundred per cent of designed life. Similarly, approximately eleven per cent of the lamps will still be burning at 150 per cent of designed life and the last lamp will have failed at slightly more than 200 per cent of designed life. The average life of the entire group of lamps, however, will be very close to 100 per cent of designed life.

Occasionally, radio manufacturers engage in lamp tests of their own, particularly with reference to checking the current rating of the lamps. Such tests have frequently been found to be wrong due to the use of low resistance voltmeters which act as a shunt around the lamp, taking from 0.02 to as much as 0.05 ampere. The ammeter then measures the sum of the lamp current and the voltmeter current so that in the case of a 6.3-volt, 0.15-ampere lamp, the reading may be anywhere from 0.17 to 0.20 ampere. It is well to note in this respect that a voltmeter having a resistance of at least 1500 ohms is required for a reasonably accurate measurement of these small currents and even with such a voltmeter the degree of error will be 0.004 ampere in the case of the lamp just mentioned.

Recommendations

A few simple precautions which have been determined through experience acquired over a number of years are to use the right lamp for the right application and in such special cases as tuning meter devices, it is also important to stress the need of making lamp replacements with the same type of lamp.

Finally, a most important item is one which has to do with making lamps accessible in receivers. In a number of models, it is necessary to remove the entire chassis from the cabinet in order to replace a burned-out lamp and since special tools are sometimes required to remove the chassis, from two to three hours is often required for this work. More often than not, the set is out of operation pending the visit of a repair man and it is only natural for the user to blame the receiver manufacturer for such short-sightedness in design.

It is important to remember that while the average life of any given group of lamps is a known factor, the life of any particular lamp is unpredictable. Since greater dependence is being placed upon panel lamps for the proper operation of modern receivers, it is only reasonable to make provisions for the easy replacement of such lamps by the users.

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A FUNDAMENTAL SUPPRESSION TYPE HARMONIC ANALYZER*

By
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Summary—*The principles and design of an harmonic analyzer of novel form are described. The analyzer operates by suppressing the fundamental and passing the harmonics through an amplifier with a calibrated gain control to a cathode-ray oscillograph or vacuum tube voltmeter.*

The apparatus is designed for analysis of audio-frequency oscillations but could be used with frequencies of a much higher order if the curve of frequency against gain were plotted for the amplifier. The power required for operation is small, an input impedance of the analyzer being of the order of 500,000 ohms.

In practice it was found that the instrument was capable of giving a fairly accurate estimate of the total harmonic content for values as low as 0.2 per cent. Using a cathode-ray oscillograph it is possible to separate two or more harmonics, but the instrument is chiefly useful for estimating total harmonic distortion of a wave.

The instrument should be useful as a distortion meter for power supply systems, giving a reading of the total harmonic distortion with the aid of a vacuum tube voltmeter.

I. Introduction

The design of a harmonic analyzer using an oscillator to beat with each harmonic and a highly selective amplifier to select and amplify the beat note, presents considerable difficulties when the voltage to be analyzed has a low audio frequency. The alternative appears to be some form of fundamental suppression method and the apparatus described below operates on this principle.

Resonance methods of analysis have been known and used for a considerable period, but they usually have the disadvantage of requiring an excessive amount of power. Morgan's[1] method is of this type. His paper has an historical summary with bibliography up to 1932. Wagner[2] gives a method free from this objection to a certain extent and with the virtue of simplicity. The other bibliographical references give some interesting resonance methods of analysis.

The method to be described has been thoroughly tested and found very effective. It is very simple to build and operate, has an input impedance of a high order, and will analyze voltages of less than 0.5 per cent distortion.

II. Principle of Method

The voltage to be analyzed is applied to the input transformer, T, in Fig. 1. This is a step-down transformer and its design is dealt with

* Decimal classification: R537.7. Original manuscript received by the Institute, June 5, 1935.
[1] Numbers refer to Bibliography.

later. The oscillatory voltage is then applied to a resistor, R_1, in series with a parallel tuned circuit consisting of an inductance, L, a capacity, C, and a resistance, R_2. The tuned circuit being adjusted to resonance with the fundamental of the distorted input voltage offers a high impedance to this voltage and if R_1 is small the greater part of the fundamental voltage is developed across LC while the harmonic voltages appear across R_1. The percentage of harmonic distortion in the voltage across R_1 is thus very greatly increased.

Since it is undesirable to reduce R_1 to less than some thousands of ohms because the impedance of LC to the lower harmonics will be comparable to that of R_1 if R_1 is small, we find it necessary to increase

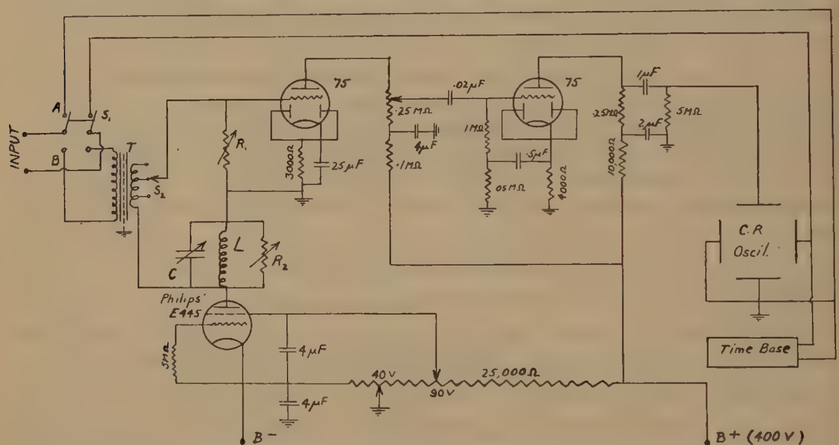

Fig. 1—Complete diagram of the harmonic analyzer, consisting of fundamental suppressor, harmonic amplifier, and analyzer equipment.

the impedance of LC to fundamental frequency. This was done in the first place by connecting LC in the grid circuit of a triode valve and using a plate coupling coil to give a variable degree of regeneration. After trying several other methods the scheme finally adopted was that shown in Fig. 1. A screen-grid tube operated as a dynatron was connected across the tuned circuit and adjusted to cause strong oscillations even when R_1 was quite small. A variable resistor, R_2, was connected across the tuned circuit to prevent oscillations, and by varying R_2 it is possible to obtain an impedance across the tuned circuit of some megohms at the fundamental frequency. If R_1 is a few thousand ohms it is thus possible to reduce the fundamental voltage relative to the harmonics in the ratio of about a thousand to one.

The voltage developed across R_1, consisting almost entirely of harmonics, is passed through an amplifier with a calibrated gain con-

trol and applied to a pair of plates of a cathode-ray oscillograph. To the other pair of plates is connected a synchronized linear time base, so that a picture of the wave form of the new voltage is obtained.

In operating the apparatus, the resistance, R_2, is increased until oscillations commence. These are visible on the screen of the oscillograph and if the time base is temporarily shut off and the switch, S_1 (Fig. 1) closed into position A, it is possible to synchronize the frequency of oscillation of the dynatron with that of the input voltage. This is done by varying the capacity of the condenser, C, and if necessary, switching the inductance L, until the image on the screen of the cathode-ray oscillograph is a steady ellipse. This procedure is only necessary when the frequency of the input voltage is unknown; in other cases, the condenser C and inductance L, can be given appropriate values immediately if they are calibrated for frequency.

With the switch, S_1, in position B and R_2 turned to zero, the wave is seen on the screen of the oscillograph with the aid of the linear time base. The reading of the calibrated gain control of the amplifier is noted and also the amplitude of the pattern on the oscillograph. With the LC circuit tuned to the frequency of the income wave, R_2 is increased until the fundamental is removed. The gain control of the amplifier is now advanced until the figure on the oscillograph again assumes its original amplitude (or some measurable fraction if the harmonic content is small). The ratio of the two readings of the gain control gives the ratio of fundamental to total harmonic.

If two or more harmonics are present they can be separated by an analysis of the new wave on the oscillograph screen. If only the total harmonic content is required, as is often the case, a vacuum tube voltmeter may be substituted for the cathode-ray oscillograph.

III. Design of Apparatus

If the resistor, R_1, and resonance circuit, LC, were connected directly across the output of the oscillator or other piece of apparatus to be tested, they would constitute a harmonic filter, damping out harmonic voltages to a certain extent. For this reason the transformer, T, is interposed. The primary of this transformer has a large value of inductance to permit of use with apparatus of large output impedance. Several secondary tappings are provided, the ratios of primary to secondary found most useful being 10/1, 3/1, and 1/1. For reasons given below, the primary and secondary should be electrostatically shielded.

The value of R_1 is found by trial, being as low as possible while still permitting of oscillation of the dynatron circuit. Values between

5000 and 10,000 ohms are usually satisfactory. The best arrangement is to make R_1 semivariable, setting it at the optimum value, found by test.

If the analyzer is to be used on audio oscillations of any frequency then the values of L and C must be variable over the necessary range. The ratio of C/L must not be large or the circuit will not offer sufficient impedance. Values of C/L of the order of one microfarad per henry and smaller were used. The resistor, R_2, may be a 0-.to 100,000-ohm variable resistor. Its effect on resonance frequency of the LC circuit is small.

The design of the amplifier portion of the analyzer (Fig. 1) requires considerable care. It follows lines similar to those of a preamplifier for microphone work, having a gain of about seventy decibels and a fairly flat response curve. It is, however, capable of handling an input of about 0.5 volt and giving undistorted output up to about sixty volts peak to a 0.5-megohm load despite the fact that type 75 tubes are used. The gain control is tapered to permit of accurate readings at low gain settings.

In operation the amplifier is made to give an output sufficient to give full deflection on the cathode-ray oscillograph. This is of the order of twenty to forty volts peak if a fairly low plate voltage is used on the oscillograph. The input is about 0.5 volt peak and when the fundamental is suppressed this may fall as low as 10^{-3} volts. If the full gain of about 3000 is now used the output voltage is three volts peak which is just enough to measure. This, of course, is the extreme case when the harmonic content is 0.2 per cent; if the harmonic exceeds this figure a more accurate measure can be made.

If the amplifier is to be operated from the alternating-current mains great care must be exercised in eliminating alternating-current ripple. It is for this reason that the input transformer, T, should have its two windings electrostatically shielded from one another. In addition to elaborate filtering of the rectified high tension voltage it was found necessary to shield all parts and wires in the grid circuit of the first amplifying tube and effectively to decouple the plate circuit of the same tube. The high tension supply voltage used is about 400 volts, this being necessary to permit of the amplifier input of 0.5 volt and output of about sixty volts without distortion.

The linear time base must be of a type which can be synchronized to any audio-frequency input voltage, unless the analyzer is to be used only on special frequencies. If it is not necessary to view the form of the composite harmonic wave then a vacuum tube voltmeter or other suitable indicating device may be used in place of the cathode-

ray oscillograph and time base. The disadvantages of using such a method of measurement are the inability to adjust quickly the resonance frequency of the LC circuit to the frequency of the input voltage, and uncertainty as to whether the fundamental has been entirely eliminated. The latter objection is removed if a preliminary test is made using the oscillograph.

IV. Operating Tests

The wave form of the output voltage of a commercial low-frequency oscillator was examined with the aid of the analyzer, cathode-ray oscillograph, and time base. The distortion, consisting almost en-

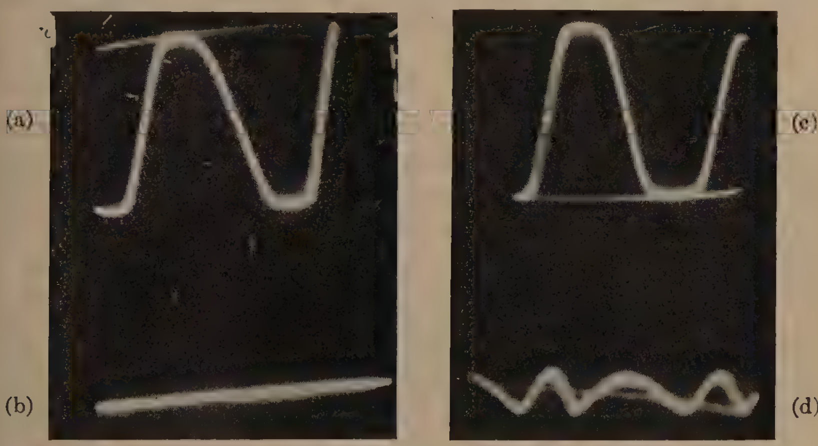

Fig. 2—Oscillograph records of low-frequency oscillations before, (a) and (c), and after, (b) and (d), suppression of the fundamental.

tirely of second harmonic could be varied by altering the degree of coupling in the oscillator circuit. For percentages of harmonic as low as 0.3 the fundamental could be reduced to an amount small compared to the harmonic. The accuracy of determination of percentages of harmonic distortion lower than this was not of a high order, as the fundamental voltage could not be reduced below about 0.1 per cent of its initial value.

It will be seen that if the C/L value of the resonance circuit is one microfarad per henry and the value of R_1 is 10,000 ohms, then the proportion of second harmonic voltage developed across the LC circuit is less than ten per cent. If necessary, correction may be made for this error.

The photographs of Fig. 2 illustrate the type of figures to be expected on the screen of the oscillograph. The first one, (a), shows a sine

wave of rather low harmonic distortion, although it appears rather distorted due to nonlinearity of the oscillograph and amplifier. The second photograph, (b), shows the residual harmonic when the fundamental has been suppressed. The gain of the amplifier has been advanced to five times its former value but the time base refuses to synchronize itself to the new composite wave. Estimating from the photographs it would appear that the harmonic distortion was of the order of two per cent.

The third photograph, (c), depicts an oscillatory voltage of very bad wave form. On suppressing its fundamental the wave of Fig. 2(d) is obtained without any amplification. We find by comparing the relative amplitudes that the voltage under test has about twenty per cent harmonic distortion, chiefly second harmonic.

Acknowledgment

The author is indebted to the Radio Research Board of the Council of Scientific and Industrial Research of Australia for rendering possible the preparation of the above paper. Also to Mr. L. T. Tindall, B.Sc., for assistance with the experimental work and to Professor J. P. V. Madsen, D.Sc., for his continued interest and advice.

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FREQUENCY CONTROL BY LOW POWER FACTOR LINE CIRCUITS*

By

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Summary—*This paper points out the advantages of concentric conductor lines as low power factor or high Q resonant circuits for controlling the frequency of very high frequency oscillators. The electrical characteristics of lines of various dimensions at various frequencies is given. Several forms of temperature compensated lines are described. Oscillator circuits, circuit combinations, and precautions for obtaining stable transmitter frequencies are suggested. Photographs of typical line controlled transmitters are included. The results obtained with line control indicate that the method has great potential usefulness comparable with the usefulness of piezoelectric crystal control.*

Introduction

IN A previous paper[1] it was pointed out that sections of transmission line could be used to stabilize the frequencies of transmitters. It was suggested that the lines, serving as very low power factor circuits, might be used advantageously to replace piezoelectric crystals in transmitters operated with relatively high power output and at very high frequencies. Since publication of the previous paper much progress has been made in the design of lines for frequency control and in their application to transmitters operated with output frequencies ranging from about seven to 500 megacycles.

The form of line best suited for frequency control is one made up of two concentric conductors, with the outer conductor completely enclosing the inner one. This form of line is relatively easy to construct and is completely shielded. The power factor of the line as a resonant circuit is not increased by radiation or coupling to surrounding objects and circuits. If desired, the outer conductor may be utilized as a means of mounting and support for tubes and other circuit elements. Copper is one of the most satisfactory materials from which to construct the line for ordinary applications but aluminum or aluminum alloys may be used where weight is an important consideration.

The degree to which a line can be made the predominating element in determining the frequency of an oscillator is proportional to the amount of oscillatory energy which may be maintained in it with a

* Decimal classification: R355.6. Original manuscript received by the Institute, July 22, 1935. Presented before Tenth Annual Convention, Detroit, Mich., July 1, 1935.

[1] Conklin, Finch, and Hansell, "New methods of frequency control employing long lines," Proc. I.R.E., vol. 10, pp. 1918–1930; November, (1931).

given amount of power. Therefore the quality or figure of merit for the line may be taken as the ratio of oscillatory energy to power loss. This ratio is popularly known as the the Q of the line.

The effective length of line may be near any number of quarter wave lengths. Making the line longer than a quarter or a half wave will not improve its Q, or sharpness of tuning, but it will increase the amount of oscillatory energy which can be stored in the line without flashover or excessive temperature rise. Usually it is preferable to obtain the desired storage rating by using large diameters rather than a length exceeding a quarter or half wave.

Characteristics of Lines

The mathematical determination of the characteristics of concentric conductor lines gives the results listed in the following table of formulas, the derivation of some of which will be given in an appendix:

Symbols

a = radius of outside surface of inner conductor in centimeters
b = radius of inside surface of outer conductor in centimeters
f = frequency
I = current in line at point of maximum current
λ = wavelength in meters

Formulas

Inductance, $L = 2\times(10)^{-7} \log_\epsilon b/a$ henrys per meter

Capacity, $C = (10)^{-9}/(18 \log_\epsilon b/a)$ farads per meter.

Characteristic impedance, $Z = 60 \log_\epsilon b/a$ ohms.

Resistance, $R = 41.6(10)^{-7}\times\sqrt{f}(1/a+1/b)$ ohms per meter for a line constructed of copper.

Attenuation constant $\alpha = R/2Z$

Power loss in a tuned line, $W = I^2R\lambda/8$ watts per quarter wave of line for a line constructed of copper.

Oscillatory energy, $VA = \pi fLI^2\lambda/4 = I^2\lambda/16\pi fC$ per quarter wave.

Figure of merit of a tuned copper line, $Q = VA/W = 2\pi fL/R = 1/2\pi fCR$.

Maximum figure of merit obtainable with a copper line, for a given value of b, is $Q_{\max} = 1/(6.86(10)^{-4}\sqrt{\lambda}/b) = 1460\, b/\sqrt{\lambda}$.

Ratio b/a giving maximum Q, for a given value of b is 3.6. (See footnotes 2, 3, and 4.)

[2] C. S. Franklin, British Patent No. 284,005 and corresponding U.S. Patent No. 1,937,559.

[3] Sterba and Feldman, "Transmission lines for short-wave radio systems," Proc. I.R.E., vol. 20, pp. 1163–1202; July, (1932).

[4] F. E. Terman, "Resonant lines in radio circuits," *Elec. Eng.*, vol. 53 pp. 1046–1061; July, (1934).

Value of maximum voltage gradient $= E/(a \log_\epsilon b/a)$.

Ratio b/a giving smallest maximum voltage gradient for a given maximum voltage and a given value of b is 2.72.

Ratio b/a giving minimum voltage gradient for a given oscillatory energy and a given value of b is 1.65.

Assuming equal thickness of inside and outside conductor the greatest oscillatory energy storage is obtainable per pound of copper when $b/a = 4.68$.

To obtain maximum impedance with a section of line, having a given value of b, (for example, for use as the equivalent of an insulator) the ratio of b/a should be 9.18.

Maxims

There are certain maxims made apparent by the study of characteristics of concentric conductor lines which are very useful in rapid interpolation of line characteristics.

1. The Q of a line is inversely proportional to the square root of the resistivity of the material used in it.
2. The Q of a line is proportional to the square root of the frequency and inversely proportional to the square root of the wavelength.
3. The Q of a line is propotional to the diameter of the conductors so long as the ratio of diameters is constant.
4. The maximum allowable oscillatory energy in a line is substantially proportional to the square of the diameters so long as the ratio of diameters is constant.[5]

Examples

Fig. 1 shows the value of the figure of merit, Q for various frequencies and diameters of the outer conductor of concentric conductor copper lines, assuming a ratio of diameters of 3.6. The values of Q for other materials and frequencies may be determined readily with the aid of maxims 1 and 2.

At 60 megacycles the minimum length of inner conductor to tune would be about 125 centimeters (49 inches). From mechanical considerations a reasonable diameter of outer conductor might be taken as 60 centimeters (24 inches). The inner conductor would be 6.5 inches in diameter. The over-all dimensions required for the finished line would be about $24 \times 24 \times 72$ inches. A line of this size would have a Q of about

[5] For more exact laws of variation in flashover voltage gradient and energy storage of lines of different dimensions see "Dielectric Phenomena in High Voltage Engineering," published by McGraw-Hill, and other publications of F. W. Peek, Jr.

20,000. Only ten watts of input power would be required to maintain an oscillatory energy of 200 kilovolt-amperes in this line.

Temperature Coefficient

Tests and theory indicate that lines made up of straight tubular conductors have a temperature coefficient of frequency variation corresponding fairly closely to the mechanical temperature coefficient of linear expansion for the material of which the line is made. So long as both conductors have the same temperature, the ratio of their diameters and, therefore, the electrical constants per unit of length do not

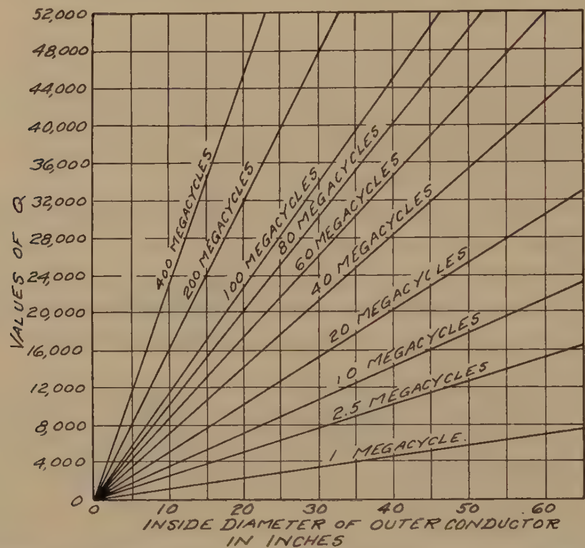


Fig. 1—Values of Q (inverse of power factor) for concentric conductor copper lines of various diameters at various frequencies, assuming a ratio of diameters of 3.6.

change with temperature. To a reasonable degree the change in frequency with change in temperature can be considered as due only to change in length.

In practice, lines used for frequency control are also subject to frequency variations due to unequal heating of inner and outer conductors. This effect is most evident in lines used with relatively large power dissipation and low frequencies and causes a temporary frequency drift while the line is warming up. The effect can be made small by using large dimensions and heavy material of good heat conductivity in the line. It is not very important in oscillators operated at 50,000 kilocycles or higher with power levels obtainable from commercially available tubes.

The approximate temperature coefficient of linear expansion, the resistivities and heat conductivities for materials which may be used advantageously in the construction of lines are as follows:

Material	Temperature Coefficient Parts per Million	Electrical Resistivity Ohm/cm$^3\times(10)^6$	Heat Conductivity Calories/cm^3/°C
Copper	16.8	1.7	0.9
Aluminum	23.1–25.5	2.8	0.5
Brass	19	6.4–8.4	0.2–0.26
Invar (1st category)	0.8 or less	80	0.025

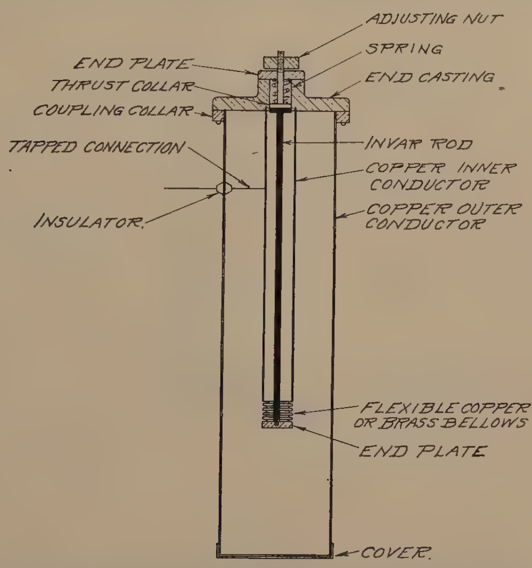


Fig. 2—Cross section of concentric conductor line one-quarter wave long having frequency adjustment and temperature compensation by means of flexible metal bellows and invar rod.

Reducing the Temperature Coefficient

One of the simplest and most effective means for reducing the temperature coefficient of frequency variation is to hold the effective length of the inner conductor constant regardless of temperature. This can be done in the manner shown in Fig. 2, where a small portion of the inner conductor is made in the form of a flexible metal bellows, and the inner conductor, including the bellows, is held constant in length by means of a rod of material, such as invar, which has a very low expansion coefficient. This construction is also particularly well adapted to making exact adjustments of frequency by adjusting the free length of the invar rod to stretch or compress the flexible bellows.

Fig. 3 shows a form of line constructed with two sizes of inner conductor in such a way that the over-all length of line required to tune to a given frequency is greatly reduced. The lengths of each of the two sizes of conductor are made substantially equal and the over-all length of both is held constant with the invar rod and flexible bellows system.

In such a line the smaller conductor and outer pipe form an effective inductance while the larger conductor and outer pipe form an effective capacity. The inductance and capacity are each very nearly proportional to the length of the respective conductors. Since the over-all length of the two inner conductors is constant, and the two are equal

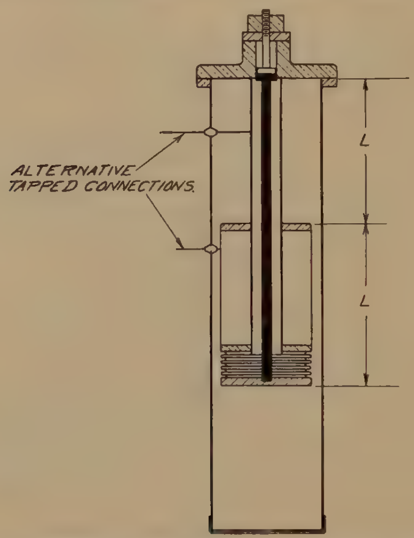
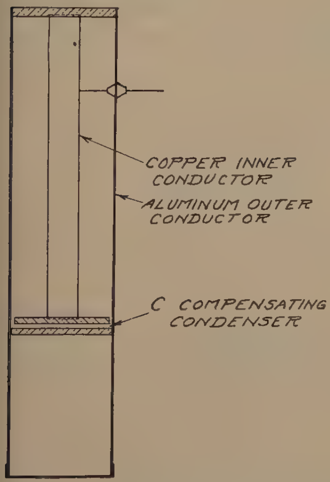


Fig. 3—Cross section of concentric conductor line having frequency adjustment and temperature compensation combined with shortening of over-all length by use of two diameters of inner conductor.

in length, any elongation or contraction of the smaller conductor, due to change in temperature, causes an equal and opposite percentage change in the larger conductor. Thus changes in temperature vary the inductance and capacity of the circuit equally and oppositely and there is little if any change in natural frequency. If lines are to be used which are physically shorter than a quarter wave, the general arrangement of Fig. 3 is a satisfactory form of construction.

Another method for reducing the temperature coefficient and at the same time shortening the line is indicated in Fig. 4. This form of line makes use of the difference in temperature coefficient of expansion of copper and aluminum to vary the capacity of C in a direction tending to compensate for variation in length of the inner conductor. If

the line temperature rises the aluminum expands more than the copper and so increases the spacing of the plates at C. This decreases the capacity at C and tends to increase the resonant frequency of the line, compensating the tendency for the frequency to decrease as the copper inner conductor increases in length. This arrangement also tends to compensate for the higher temperature rise of the inner conductor due to losses in the line. Fig. 5 is an inverted variation of the arrangement shown in Fig. 4. Both these arrangements require careful design, construction, and adjustment and are not so convenient to use as the lines of Figs. 2 and 3.



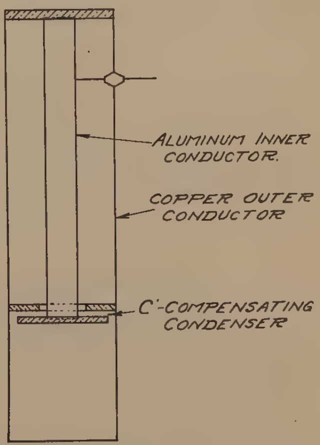


Fig. 4—Cross section of temperature compensated line employing copper inner conductor and aluminum outer conductor.
Fig. 5—Cross section of temperature compensated line employing aluminum inner conductor and copper outer conductor.

Figs. 2, 3, 4, and 5 show quarter-wave lines which are most obviously applicable to single tube oscillators. Of course, the lines may also be made a half-wave long, equivalent to two quarter-wave sections in series, in which case they are most readily applicable to push-pull oscillators. To save over-all length the half-wave system may be bent into the shape of a U as shown in Fig. 6.

Fig. 6 shows a means for compensating the effects of temperature variation by using the relatively large variation in volume of some liquids, such as lubricating oil, with change in temperature to operate a condenser plate. The oil is contained in a pipe system exposed to the same temperature variations as the line and operates the condenser plate through a flexible metal bellows and a lever arm. By suitably locating the oil piping and varying its insulation, together with proper

adjustment of the condenser plate and bellows it is possible to obtain quite close compensation for temperature variations due to changes in ambient temperature and due to line losses. However, considerable skill and patience is required in making the design and adjustment.

Eventually, when materials of low temperature coefficient become more readily available in commercial shapes it is probable that they will be used to obtain lines having low temperature coefficients without compensating arrangements. Lines constructed of these materials will require silver or copper plating of the conducting surfaces.

Internal stresses in the material of the lines, which may cause progressive changes in dimensions due to temperature cycling, should be avoided by care in manufacturing or by suitable heat treatment. This

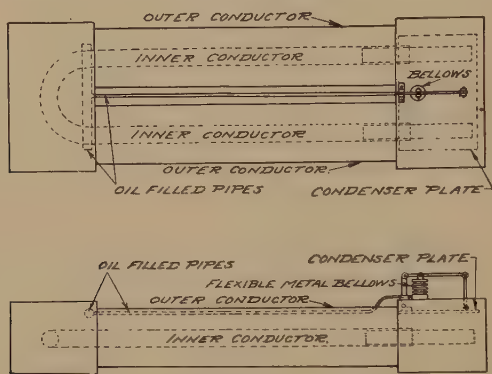


Fig. 6—Half-wave line with temperature compensation by means of condenser plate controlled by expansion and contraction of a liquid.

source of frequency variation is most troublesome in newly constructed lines and will usually decrease with age.

Circuits for Line Controlled Oscillators

In general, lines may be used for frequency control, and will operate in a manner similar to piezoelectric crystals. They differ from crystals chiefly in their ability to control high power oscillators and in their ability to perform satisfactorily at frequencies far higher than can be reached by crystals. As we go to higher frequencies, crystals gradually become less useful in stabilizing oscillators while lines increase in their stabilizing ability.

Fig. 7 shows a line controlled oscillator similar to one often used for crystal oscillators. In operating this circuit it is best to adjust the regeneration control condenser to give about the smallest feedback from plate to grid circuit which can be used to make the oscillator work efficiently. Any excess feedback reduces the ability of the line to sta-

bilize the frequency. The circuit will function with the regeneration control condenser set either above or below the capacity value required for a balance but one adjustment or the other will be preferable depending upon the ratio of effective resistance in anode and grid circuits and the frequency. The resistance ratio, electron time lag at very high frequencies and feed-back adjustment are all factors which should be taken into account to determine which adjustment will give the best phase relation between anode and grid radio-frequency voltages.

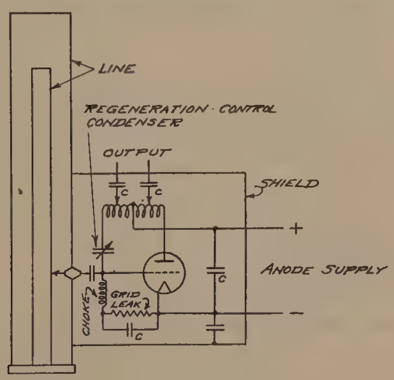


Fig. 7—Single tube line controlled oscillator circuit.

Fig. 8 is a push-pull oscillator circuit similar in principle and operation to the single tube circuit of Fig. 7.

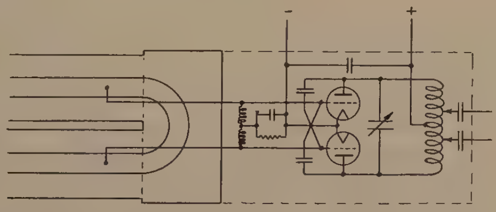


Fig. 8—Push-pull line controlled oscillator circuit with half-wave line.

Fig. 9 is a circuit suitable for stabilizing the frequency of a push-pull oscillator by means of a quarter-wave line. In this circuit the grids of the two tubes are inductively coupled to the line by means of coupling loops of opposite polarity.

Line Controlled Transmitter Combinations

In some cases, where the antenna system is made mechanically rigid and free from variations in input impedance due to weather, it is possible to obtain acceptable frequency stability with the oscillator

coupled directly to the antenna. For very high frequency transmitters of limited range, in places where interference is not a problem, single stage transmitters will often be entirely satisfactory and can be recommended. The interisland radiotelephone system of the Mutual Telephone Company, in Hawaii, where single stage line controlled transmitters have been in regular commercial operation since 1931, is an excellent example.[1,6]

In most cases, however, it is desirable or necessary to interpose one or more stages of amplifier between the line controlled oscillator and

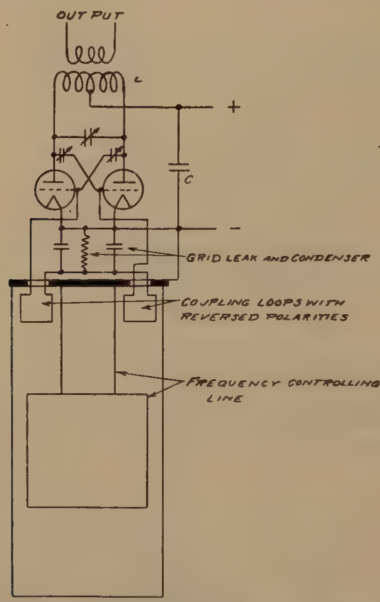


Fig. 9—Push-pull line controlled oscillator circuit with quarter-wave line and inductive coupling to line.

the antenna. For most ordinary requirements a single radio-frequency amplifier between oscillator and antenna will suffice if the amplifier is carefully neutralized and shielded to prevent feedback to the oscillator.

For very accurate frequency control, such as will be required in the future to make maximum use of very high frequencies, two stages of amplifier following the oscillator are recommended. At frequencies near the upper limit for the tubes it is desirable to operate the master

[6] Beverage, Peterson, and Hansell, "Application of frequencies above 30,000 kilocycles to communication problems," Proc. I.R.E., vol. 19, pp. 1313–1333; August, (1931).

oscillator at a half or a third of the final output frequency and to follow it with a frequency multiplying amplifier and a power amplifier. The use of frequency multiplication permits the oscillator to work at a lower frequency where the tubes are more efficient and very greatly reduces the probability of frequency changes due to variable radio-frequency feedback from the later stages.[7]

When line control is used to control a relatively low-frequency transmitter it will sometimes be desirable to operate the oscillator at two or three times the output frequency and to follow it with a controlled oscillator and power amplifier, both operated at the output frequency. This greatly reduces the dimensions of line, and improves its frequency holding ability. By careful design and adjustment the controlled oscillator can be made to serve as an uncoupling link between the oscillator and power amplifier with an effectiveness about equal to that obtained from a frequency multiplier.

The controlled oscillator should have weak regeneration at the output frequency so that the grid voltage at the output frequency and the grid voltage at the harmonic input frequency will not be greatly different. It may also be noted that the controlled oscillator should not be amplitude modulated by any great amount. Amplitude modulation should be applied only to the power amplifier.

Precautions for Reducing Undesired Modulations

In general, line controlled oscillators require the same precautions for obtaining a pure continuous wave output as are required for crystal oscillators. Ripples in the direct power voltages and alternating cathode heating current all tend to produce undesired amplitude phase, and frequency modulations of the output. Because of the line these undesired modulations will be far less than they would be in a simple oscillator but they will always be present.

The most obvious means for reducing these modulations are to use very smooth direct voltages and direct-current cathode heating. Both of these expedients are undesirable from the standpoint of cost, simplicity, and reliability but may be necessary in some cases. In other cases, satisfactory results may be obtained while using alternating-current cathode heating and impure direct-current anode supply by taking a few simple precautions.

Amplitude modulations introduced in all but the last stage, may be kept small by using sufficient excitation in the later stages to produce limiting. If the last stage is amplitude modulated by means of the

[7] Hallborg, Briggs, and Hansell, "Short-wave commercial long-distance communication," Proc. I.R.E., vol. 15, pp. 467–500; June, (1927).

Heising constant current modulating system this stage is automatically provided with very good anode supply smoothing by the modulation choke.

Elimination of undesired phase and frequency modulations requires holding constant tube impedances and the use of circuits tending to minimize the effect of changing impedances upon the tuning of the circuits. Small grid current and the use of grid leak and cathode return resistor biasing, particularly in the oscillator and next succeeding amplifier, assist materially in holding constant effective grid impedances, provided the resistances have sufficiently small parallel dielectric capacity to prevent appreciable phase lag in bias variation in response to radio-frequency amplitude variations. Also the effective series radio-frequency reactance from the anodes of one stage to the grids of a succeeding stage must be made a minimum. Leads from the output circuit of one stage to the input circuit or grids of a succeeding stage should be large and extremely short or else equal to a half wave or multiples of a half wave long. Leads which are near a quarter wave or multiples of a quarter wave in length should be carefully avoided.

It is interesting to note that any phase or frequency modulation noise introduced in an early stage of a transmitter will be increased in proportion to the amount of frequency multiplication used after that stage. A crystal controlled transmitter having an output of 100,000 kilocycles would probably start out with an oscillator frequency of about 3125 kilocycles. One degree of phase modulation in the output of the crystal oscillator would then appear as thirty-two degrees in the output of the transmitter and produce side frequency energy equivalent to that obtained with about sixty per cent amplitude modulation. With line control the oscillator may be operated at the output frequency so that one degree of phase modulation in the oscillator will appear as one degree in the transmitter output.

Mechanical vibration of the line and circuits must be prevented. This requires rigid construction of the line and all coils, condensers, leads, etc. Poor workmanship must be avoided. If the equipment is to be operated near rotating machinery or other sources of vibration it is desirable that the whole radio-frequency system be hung on springs and rubber shock absorbers so adjusted that the rubber is subjected to little initial stress.

Examples of Line Controlled Transmitters

Transmitters WQO and WHR

Figs. 10 and 11 show the general construction and mounting of the lines used to control the frequencies of transmitters WQO—6725

kilocycles and WHR—13,420 kilocycles, respectively, at the Rocky Point, New York, station. These two transmitters each have two RCA-846 tubes in a push-pull master oscillator, employing the circuit

Fig. 10—Line controlled master oscillator for commercial transmitter WQO-6725 kilocycles.

Fig. 11—Line controlled master oscillator for commercial transmitter WHR-13,420 kilocycles.

shown in Fig. 8, and two UV-858 tubes in a push-pull power amplifier. The master oscillators are located inside the boxes at the ends of the lines, above the power amplifier units. Keying is accomplished by grid-bias control of the master oscillators. Both transmitters are supplied

with anode power from a common rectifier. The two transmitters are operated alternately for day and night service but can be operated simultaneously at reduced power. Both are started, stopped, and keyed by remote control over the control circuit from New York, about sixty miles away. Application of keying energy to the control circuits in New York automatically starts up either transmitter and removal of the keying for several minutes automatically shuts it down.

These two line installations are each a half wave long and employ the arrangement illustrated in Fig. 6 for compensating the effect of temperature variations upon the frequency. The line for WQO—6725 kilocycles has an outer pipe eight inches inside diameter, #10 Stubs gauge copper and an inner pipe of two-inch iron pipe size copper. The line is built up in four sections joined together at the ends into one half-wave oscillator circuit.

The line for WHR—13,420 kilocycles has an outer pipe twenty inches diameter, #12 Stubs gauge copper and an inner pipe of five-inch iron pipe size copper. It is built up in two sections bent into one U-shaped oscillator circuit.

The RCA Central Frequency Bureau, at the Riverhead, New York, receiving station, makes frequent routine checks on the frequencies of all transmitters operated by R.C.A. Communications, Inc., as well as those of most other transmitters engaged in long-distance service. These routine checks have been utilized to determine the relative effectiveness of line control in maintaining the frequencies of WQO and WHR.

Taking the period of May 1, 1934, to April 30, 1935, as representative, these checks show maximum variations of WQO and WHR to be 0.01 and 0.015 per cent, respectively. The average of maximum variations reported in any one week were WQO—0.0052 and WHR—0.0056 per cent. The average of weekly maximum variations reported for a representative group of fifty crystal controlled transmitters in the same period was 0.0115 per cent.

Fig 12 is a photograph of W2XHG—25,700 kilocycles, 100 to 150 watts, located on the roof of the RCA Building, New York City. This transmitter as shown in the photograph had a line controlled master oscillator followed by a modulated power amplifier, utilizing RCA-852 tubes. The line was constructed in the manner illustrated in Fig. 3. For propagation survey purposes the transmitter has been adjusted successively for a range of frequencies from 25,700 kilocycles upwards. The antennas usually used have been horizontal dipoles at the top of a mast which extends from the transmitter room to about twenty-eight

feet above the roof. The base of the mast may be seen at the left of the transmitter. The transmission line to the antennas was run inside the mast.

Fig. 12—Line controlled transmitter on roof of RCA Building, 30 Rockefeller Plaza, New York City, operated at 25.7 megacycles or higher.

Fig. 13 is a photograph of W2XBN—91,800 kilocycles, 100 watts, located on the top floor of the Continental Bank Building at 30 Broad Street, New York City. This transmitter utilizes RCA-852 tubes and has a line controlled oscillator, similar to Figs. 2 and 7, at 45,900 kilocycles followed by a frequency doubler and an amplifier system. It provides a multiplex radio control circuit between the central traffic

office of R.C.A. Communications, Inc., at 64 Broad Street, New York City, and the transoceanic station at New Brunswick, N.J. A directive antenna on the roof provides a power gain of about eight to one.

Fig. 14 is a photograph of W2XS—200,000 kilocycles designed with a master oscillator utilizing RCA-800 or RCA-834 tubes followed by a power amplifier utilizing small water-cooled tubes. Both stages operate at the same frequency. The final output is about 250 watts.

Up to the time of writing this paper about twenty-five or thirty line controlled transmitters for both experimental and commercial

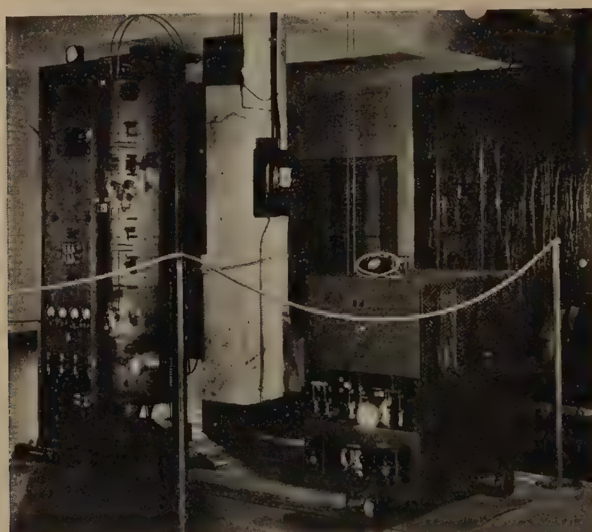


Fig. 13—Line controlled transmitter on roof of Continental Bank Building, 30 Broad Street, New York City, operated at 91.8 megacycles.

service have been built. At the lowest frequency, 6725 kilocycles, the power output was about thirty kilowatts. At the highest frequency, 450,000 kilocycles, the power output has ranged up to 110 watts.

Conclusion

Our experience during the past six years with line control of oscillator frequencies indicates that the method has great potential usefulness. It seems almost certain that it will provide a frequency stabilizing device for use at frequencies above about 20,000 kilocycles of as great practical value as piezoelectric crystals have been for use at lower frequencies.

Appendix

Radio-Frequency Resistance

In order to illustrate the principles involved in the phenomena of skin effect we shall first consider a very simple case. Assume a tubular line in which the inside conductor consists of two thin tubes a and b, differing in radius by da connected in parallel and in which the single tube outside conductor has the radius C. (See Fig. 15.)

Fig. 14—Development model line controlled transmitter for operation at 200 megacycles.

Faraday's law states that the line integral of electric force around any closed circuit is equal to the negative rate of change of magnetic flux through the circuit, in accordance with the right-hand screw rule. Taking a path of unit length on conductors a and b in a plane cutting them longitudinally, we have $E_a - E_b = d/dt(\Phi_{ab})$. The magnetic force H between a and b is equal to $2i_a/a$. If the current is sinusoidal and represented by the real part of $I_a \epsilon^{j\omega t}$, $d/dt(i_a) = j\omega i_a$ and $E_a - E_b = -j\omega(2i_a/a)da$, as the area of a unit length is da. If R is the resistance of both a and b we have from Ohm's law:

$$E_a = i_a R \text{ and } E_b = i_b R.$$

Hence $R(i_a - i_b) = -i_a(2j\omega(da/a))$. $2da/a$ is the inductance per unit length of a circuit consisting of tubes a and b alone. Calling this L_{ab} we have:

$$R(i_a - i_b) = -j\omega L_{ab} i_a$$

or,

$$i_a/i_b = \frac{R}{R + j\omega L_{ab}}.$$

From this relation it is apparent that if either the resistance approaches zero or the frequency approaches infinity, the current will all flow in the tube of larger radius. Also it should be noted that the radius

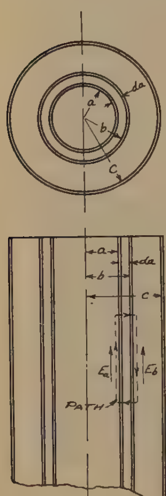


Fig. 15

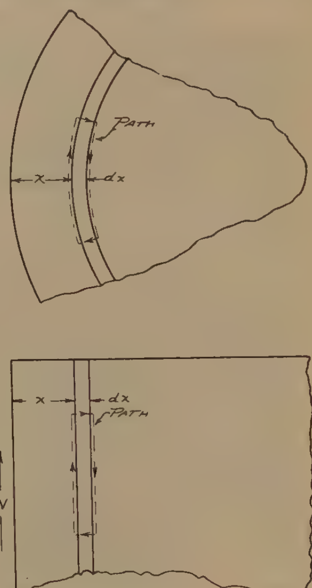


Fig. 16

C of the return conductor has no effect upon the distribution of the current between a and b.

Let us now consider the actual problem of current distribution within a solid conductor. For frequencies of the order of megacycles we know the penetration of current to be very small. If we assume this at the start so that we can neglect a change in resistance of successive thin cylinders as the radius is decreased, the problem of determination of effective resistance is greatly simplified. We shall proceed upon this assumption. Call the distance from the surface of the conductor X and the distance along the axis Z. (See Fig. 16.)

Let u = the current density = the real part of $U\epsilon^{j\omega t}$ so that $du/dt = j\omega u$.

Take a cylinder of thickness dx at a distance X from the surface and draw a circuit of unit peripheral length. The total current within this circuit is udx, the magnetic force at X is H and at $X+dx$ is $H+(\delta H/\delta x)dx$. The line integral around the circuit is then $H-(H+(\delta H/\delta x)dx)=-(\delta H/\delta x)dx$.

This by Ampere's law is equal to 4π times the total current included.

Hence $4\pi udx=-(\delta H/\delta x)dx$ or $4\pi u=-\delta H/\delta x$.

Now take a circuit in a longitudinal plane. Since the current in the first case was assumed flowing into the paper the line integral of electric force around a circuit of unit length is

$$-E+\left(E+\frac{\delta E}{\delta x}dx\right)=\frac{\delta E}{\delta x}dx.$$

This by Faraday's law is equal to the negative rate of change of magnetic induction included by the circuit, or $\mu d/dt(Hdx)$.

Hence,

$$\frac{\delta E}{\delta x}dx=-\mu\frac{dH}{dt}dx \text{ where } \mu = \text{permeability}$$

or,

$$\frac{\delta E}{\delta x}=-\mu\frac{dH}{dt}.$$

Differentiating with respect to X and substituting for $\delta H/\delta x$ we get

$$\frac{\delta^2 E}{\delta x^2}=-\mu\frac{d}{dt}\left(\frac{\delta H}{\delta x}\right)=4\pi\mu\frac{du}{dt}=4\pi\mu j\omega u.$$

However, from Ohm's law $E=\rho u$ where ρ is the resistivity. Hence $\delta^2u/\delta x^2=j(4\pi\omega u/\rho)$. The solution of this differential equation is

$$u=u_0\epsilon^{-(\alpha+j\alpha)x}$$

where,

$$\alpha=\sqrt{2\pi\omega\mu/\rho}.$$

The total current per unit peripheral length is then

$$i_0=u_0\int_0^\infty \epsilon^{-\alpha(1+j)x}=\frac{u_0}{\alpha(1+j)}=\frac{u_0}{\sqrt{2}\alpha}\epsilon^{-j(\pi/4)}.$$

The average heat loss per unit area and unit length is $(U^2/2)\rho dx$ or $(U_0^2/2)(\epsilon^{-\alpha x})^2dx$.

The total heat loss per unit width is $(U_0^2\rho/2)\int_0^\infty \epsilon^{-2\alpha x}dx = U_0^2\rho/4\alpha = \rho(i_0^2/\alpha)$.

Hence the effective resistance $= \rho\sqrt{2\pi\mu\omega/\rho} = \sqrt{2\pi\mu\rho\omega}$ per unit width.

For a radius r the resistance is $R = \sqrt{2\pi\mu\rho\pi}/2\pi r = \sqrt{\mu f\rho}/r$ electromagnetic units.

In practical units $R = \sqrt{\mu f\rho/r} \times 10^{-9}$ ohms per centimeter length for r in centimeters.

For copper $\rho = 1724$ electromagnetic units, $\mu = 1$ and $R = (41.5/r)\sqrt{f} \times 10^{-9}$ ohms per centimeter of length.

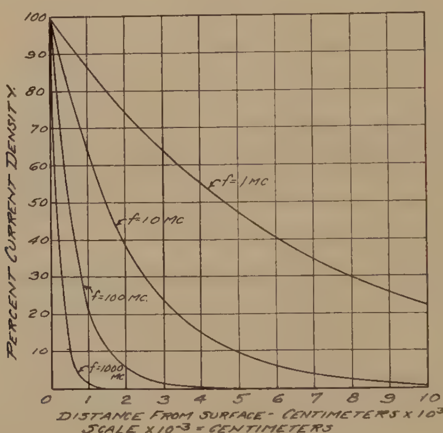


Fig. 17

The relation for current distribution shows that the current density falls off exponentially as we proceed away from the surface into the wire and the phase shifts through an angle of ninety degrees in proceeding from the conductor surface to an internal position where the value of the current is negligible. The phase angle of the total current lags that of the surface current by forty-five degrees. We may consider this phase-shift effect to be due to an internal self-inductance. It may be of interest to note the manner in which the current distributes itself near the surface of a conductor at different frequencies. This is shown for several frequencies in Fig. 17. It may be noted that for a frequency of ten megacycles the current density becomes ten per cent of its value at the surface at a distance of five thousandths of a centimeter. It is apparent from the foregoing reasoning that the same relations hold for both the internal and external conductors of a concentric tube line. Hence for the total resistance of a unit length of concentric copper tube line we have $R = 41.5 \times 10^{-9}$

$\sqrt{f}(1/a+1/b)$ ohms per centimeter where a and b are the radii of the tubes in centimeters.

Power Factor

A tuned transmission line short-circuited at the far end an odd number of quarter waves long has a distribution of voltage given approximately by $E=E_0 \cos 2\pi x/\lambda$ and a current distribution given by $i=(E_0/Z_0) \sin 2\pi x/\lambda$ where X is the distance, Z_0 the characteristic impedance $=\sqrt{L/C}$, and E_0 the input voltage. For open-circuited lines any integral number of half waves long the distribution of voltage and current is given by $E=I_0 Z_0 \sin 2\pi x/\lambda$ and $I=I_0 \cos 2\pi x/\lambda$.

The power dissipated in a section of length dx is $i^2 R dx$ where R is the resistance per centimeter.

The total power is then $W=\int_0^l I_0^2 R \sin^2(2\pi x/\lambda)dx=(1/2)I_0^2 Rl$.

Hence the effective resistance is $R/2$. The displacement current in a length dX is $E\omega c dx$ and the voltamperes $(VA)=E^2\omega c dx$ where c is the capacity per centimeter of length.

The total volt-amperage is then

$$VA = \int_0^l \omega c E_0^2 \sin \frac{2\pi x}{\lambda} dx = \frac{\omega c E_0^2 l}{2} = \frac{\pi E_0^2 l}{z_0 \lambda} = \frac{I_0 l}{\omega c}.$$

For all practical purposes the effective power factor $=\dfrac{\text{watts}}{\text{VA}}$

$$=\frac{\frac{1}{2} I_0^2 Rl}{\frac{1}{2} E_0^2 \omega c l} = \frac{R}{Z_0^2 \omega c}.$$

But $Z_0=60 \log_\epsilon b/a$

and $C=\dfrac{1}{2 \log_\epsilon b/a} \times \dfrac{1}{9} \times 10^{-11}$ farads per centimeter

and the power factor $=PF=\dfrac{R}{2\times 10^{-9}\omega \log_\epsilon b/a}$ where R is in ohms

per centimeter length but $R=\sqrt{\rho f}(1/a+1/b)\times 10^{-9}$ assuming non-magnetic conductors.

Hence $PF=\left(\dfrac{1}{a}+\dfrac{1}{b}\right) \dfrac{\sqrt{f\rho}}{2\omega \log_\epsilon b/a}$

and for copper $PF=\dfrac{3.3}{\sqrt{f}}\left(\dfrac{1}{a}+\dfrac{1}{b}\right) \dfrac{1}{\log_\epsilon b/a}$ for a and b in centimeters.

The power factor may be expressed in terms of the attenuation constant in which it becomes $PF=\alpha\lambda/\pi$ where α is taken for the same unit of length as λ.

When b/a is made the best ratio, 3.6, for minimum loss, the power factor becomes $PF=\frac{1}{4\pi a}\sqrt{\frac{\mu\rho}{f}}$ or $Q=\frac{1}{PF}=4\pi a\sqrt{\frac{f}{\mu\rho}}$ and for copper $PF=1.905\times10^{-5}\frac{\sqrt{\lambda}}{a}$, for a and λ in centimeters, and

$$Q = 5.25 \times 10^{-4}\frac{a}{\sqrt{\lambda}}.$$

Voltage Gradient

The maximum electric force or voltage gradient E for a concentric tube line is given by $E_a=\frac{V}{\log_\epsilon b/a}\times\frac{1}{a}$ volts per centimeter where V is the voltage. For a given voltage and given outside tube diameter this becomes a minimum when $b/a=\epsilon\cong2.72$. However, this is not usually of interest. The important consideration is that of obtaining a maximum of volt-amperage for a given gradient. In terms of volt-amperage the gradient E becomes

$$E = \frac{60}{\pi l/\lambda}/a(\log_\epsilon b/a)^{1/2}.$$

Minimizing this expression we find $b/a=\epsilon^{1/2}\cong1.65$.

For this ratio the gradient becomes

$$E = \frac{10.2}{b}\sqrt{\frac{VA}{l/\lambda}} \text{ volts per centimeter.}$$

When the line is designed for a minimum loss ($b/a=3.6$) the gradient is thirty-six per cent greater than the value given above.

Input Impedance

The input impedance Z_0 of a quarter-wave line short-circuited at its far end or of a half-wave line open at its far end is

$$Z_i = \frac{Z_0}{\tanh \alpha l}\cong\frac{Z_0}{\alpha l}$$

where α is the attenuation factor.

However, $\alpha = R/2Z_0$ and therefore

$$Z_i = \frac{2Z_0^2}{Rl}$$

which in terms of b and a becomes

$$Z_i = \frac{2.60^2(\log_\epsilon b/a)^2}{41.5 \times 10^{-9}\sqrt{f}(1/a + 1/b)l}$$

Maximizing this expression we find Z_i to be a maximum when $b/a = \epsilon^{2(1+a/b)}$ or $b/a = 9.18$.

When the ratio is 9.18 the input impedance becomes

$$Z_i = \frac{8.4 \times 10^{10}}{\sqrt{fl}} \times b \text{ ohms, for } b \text{ and } l \text{ in centimeters}$$

$$= 11.2b\sqrt{f} \quad \text{for a one-quarter-wave-length line (closed)}$$

or $5.6b\sqrt{f}$ for a one-half-wave-length line (open).

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CALCULATION AND DESIGN OF CLASS C AMPLIFIERS*

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Summary—*A method of calculating the performance of class C triode amplifiers is presented which is based on the assumption that the total space current* (I_p+I_g) *is proportional to* $(E_g+E_p/\mu)^\alpha$, *where* α *is a constant, usually close to three halves. The direct-current and fundamental alternating-current components of such a space current pulse are presented (graphically) as a function of the angle of current flow, for various values of* α, *and it is shown how to obtain the plate current components by estimating the direct grid current, and correcting for the current diverted to the grid.*

It is shown how the results of the analysis may be applied in a straightforward manner to lay out class C amplifiers on paper, and to predict power output, power input, plate loss, etc., for any particular set of operating conditions. The method is applied to several examples and the accuracy shown to be satisfactory for all ordinary design requirements.

In designing a class C amplifier one normally starts with a particular tube and attempts to realize the optimum operating conditions. This can be done either by setting up the amplifier and following a cut-and-try process, or by making preliminary calculations on paper. The latter method is the most satisfactory since it is much quicker, particularly when large tubes are involved, and is also more certain of yielding the optimum design.

An exact calculation of class C amplifier performance requires that a complete set of characteristic curves be available. With this information one can trace out the plate and grid-current pulses for any operating condition, following the method originally devised by D. C. Prince.[1] This procedure can be simplified somewhat by plotting the tube characteristics in the form of constant current curves as described by Mouromtseff and Kozanowski.[2] Such point-by-point calculations give the exact performance, but unfortunately the amount of labor involved is such that class C amplifiers have usually been built on the cut-and-try basis.

In the last several years a number of approximate methods of de-

* Decimal classification: R355.7. Original manuscript received by the Institute, November 18, 1935.

[1] D. C. Prince, "Vacuum tubes as power oscillators," Proc. I.R.E., vol. 11, pp. 275, 405, 527; June, August, and October, (1923).

[2] Mouromtseff and Kozanowski, "Analysis of the operation of vacuum tubes as class C amplifiers," Proc. I.R.E., vol. 23, pp. 752–778; July, (1935).

signing class C amplifiers have been described.[3,4,5,6] These are all based upon the assumption that the plate current of the tube can be represented by some simple law and that the grid current can be neglected. Some of them are also limited to such special operating requirements as a maximum positive grid voltage equal to the minimum instantaneous plate potential. As a result none of these methods of approximate analysis is entirely satisfactory for practical problems.

The purpose of the present paper is to present a means of calculating class C amplifier performance which is both simpler and more accurate than the approximate analyses mentioned, and which at the same time does not neglect the grid current and is not limited to special operating conditions. The method is essentially an extension of a procedure described in a previous paper,[7] and has been in use for several years at Stanford University.

Voltage and Grid Relations in Class C Amplifiers

The circuit and fundamental voltage and current relations of a class C amplifier are illustrated in Fig. 1. The voltage actually applied to the grid of the tube consists of the grid bias E_c plus the exciting voltage E_s. The relations are normally such that at the crest of the cycle the grid is driven appreciably positive and consequently draws some grid current. The voltage actually appearing at the plate of the tube consists of the battery voltage E_b minus the voltage drop E_L in the plate load impedance, and so has the wave shape shown in Fig. 1(a). The phase relations are such that the minimum instantaneous plate potential $E_{\min}$ occurs the same part of the cycle as the maximum grid potential $E_{\max}$. The alternating components of the plate and grid voltage are also always sinusoidal since they are developed across sharply resonant circuits.

The plate and grid currents that flow at any instant are the result of the combined action of the plate and grid potentials at that instant, and can be determined from these potentials with the aid of a set of complete characteristic curves of the tube. The plate current is in the form of an impulse flowing for something less than half a cycle. The grid cur-

[3] L. B. Hallman, Jr., "Fourier analysis of radio-frequency power amplifier wave forms," Proc. I.R.E., vol. 20, pp. 1640-1659; October, (1932).

[4] W. L. Everitt, "Optimum operating condition for class C amplifiers," Proc. I.R.E., vol. 22, pp. 152–176; February, (1934).

[5] Burton F. Miller, "Analysis of class B and class C amplifiers," Proc. I.R.E., vol. 23, pp. 496–510; May, (1935).

[6] A. P. T. Sah, "The performance characteristics of linear triode amplifiers," *Science Reports of National Tsinghua University*, Peiping, China, vol. 2, pp. 49 and 83; April and July, (1933).

[7] F. E. Terman and J. H. Ferns, "The calculation of class C amplifier and harmonic generator performance of screen-grid and similar tubes," Proc. I.R.E., vol. 22, pp. 359–373; March, (1934).

rent flows only when the grid is positive, and is usually sharply peaked. In some cases the grid current may reverse and be negative for a portion of the time as a result of secondary emission. The sum (I_p+I_g) of plate and grid currents represents the total space current flowing away from the filament, and is shown in Fig. 1(c). This current always has its peak at the instant when the grid and plate potentials are $E_{\max}$ and $E_{\min}$, respectively.

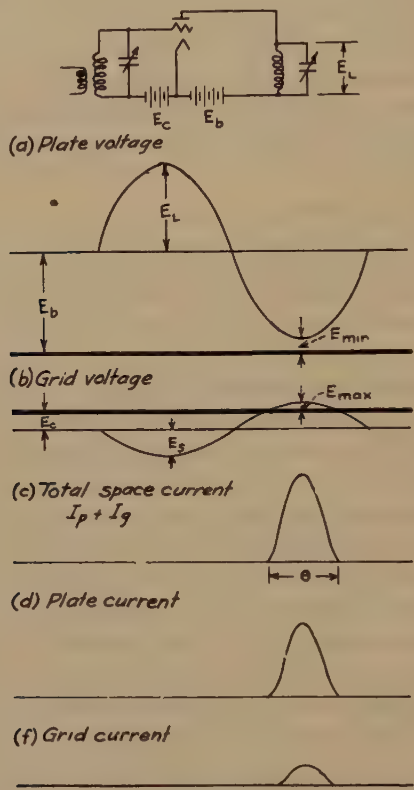

Fig. 1—Circuit and voltage and current relations of class C amplifier.

The average value of the plate current pulse over a complete cycle represents the direct current which will be observed in the plate circuit, while the average value of the grid current pulse is likewise the direct grid current. The power input which must be supplied by the plate battery is the direct plate current times the plate supply voltage E_b. The power delivered to the load is equal to half the product of alternating voltage $E_L = E_b - E_{\min}$ across the load and the crest value of the fundamental frequency component of the plate current impulse.

Basis of Analysis

The analysis given in this paper is based upon the assumption that the total space current (I_p+I_g) can be expressed by the following mathematical relation

$$\text{total space current} = (I_p + I_g) = K(E_g + E_p/\mu)^\alpha \tag{1}$$

where E_g and E_p are instantaneous plate and grid voltages, respectively, μ is the amplification factor of the tube, and K and α are constants. The exponent α in (1) is normally very close to three halves and would be exactly three halves if the tube were perfectly symmetrical and had full space-charge saturation. Actually it is found that the characteristic of ordinary power tubes follows the relationship given in (1) reasonably well.

Upon the assumption that the space current is of the form given by (1), one can plot curves which give the relationship between the direct current and fundamental frequency components of the total space current in terms of the maximum space current I_m and the number of electrical degrees θ during which the plate current flows. Such curves have been calculated according to the method given in the Appendix for various values of α between 1.0 and 2.0, and are presented in Fig. 2. The importance of these curves is that they enable one to determine the direct-current and fundamental alternating-current components of the space current impulse without the necessity of resorting to point-by-point calculations.

Design Procedure

The procedure for designing a class C amplifier making use of the curves of Fig. 2 involves a sequence of steps as outlined below.

First. The first step is the choice of a suitable value of crest space current I_m. The maximum value permissible is determined by the electron emission which the filament is capable of producing, and in order to use the full possibilities of the tube it is usually desirable to select the highest possible value of I_m. With tungsten filaments it is common practice to make I_m equal substantially the full emission from the filament in the case of class C amplifiers, and perhaps two thirds this for modulated and class B amplifiers where linearity is important. With thoriated tungsten filaments the deterioration during life is such that factors of safety of three to seven are common, with the exact value depending upon how thoroughly the tube has been evacuated. The characteristics of oxide-coated filaments vary so much that still higher factors of safety must be employed with them.

Second. After an appropriate value of maximum space current has

been determined one next selects a combination of values for maximum grid potential E_{max}, and minimum plate potential E_{min}, that will draw this total space current. What is desired is the lowest possible value of minimum plate voltage E_{min} because the lower the plate voltage the higher will be the efficiency of the amplifier. In order to draw the full space current with a low minimum space current it is then necessary to make the maximum grid potential E_{max} large. However, the maximum grid potential should never under any conditions exceed the minimum plate voltage since this will cause the grid current to be excessive.

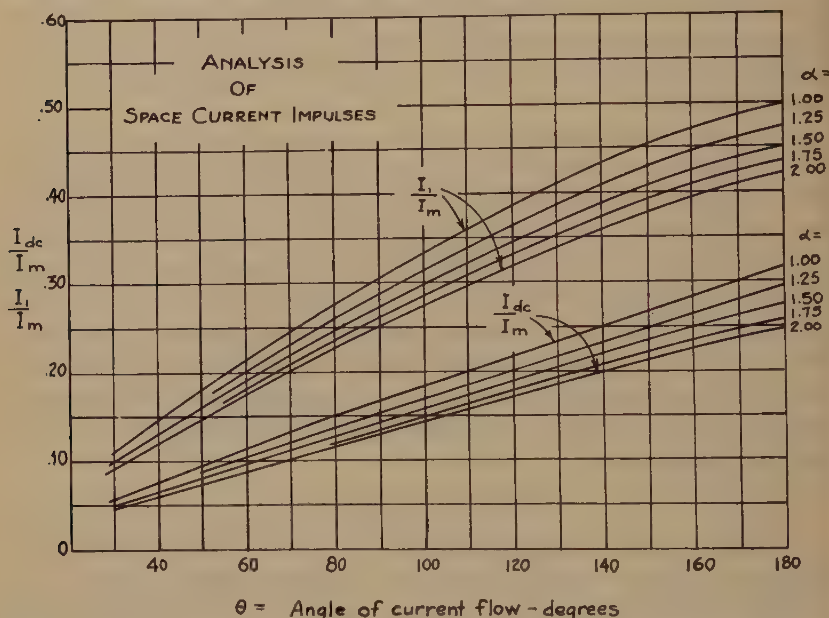

Fig. 2—Curves giving the direct-current and fundamental frequency components of the space current impulse as a function of the angle of flow, for various values of α in equation (1).

This large grid current is diverted away from the plate, thereby reducing the output power, and also represents large driving power since it means a high power consumption in the grid circuit. The usual practice is to make the maximum grid potential approximately equal to the minimum plate potential with tubes which operate at a few thousand volts plate potential, while with large water-cooled tubes the maximum grid potential is normally considerably less, perhaps one half to one fifth of the minimum plate potential.

When complete tube characteristics are available one can determine from them combinations of E_{min} and E_{max} which will draw the desired

value of total space current. When complete characteristics are not available, satisfactory results can be obtained by plotting the total space current (I_p+I_g) as a function of effective anode voltage (E_g+E_p/μ) on log-log paper, covering the range of values which can be taken in the usual point-by-point way without overheating the tube. This curve will be substantially a straight line and can then be extrapolated to the desired total space current I_m.

Third. One is now ready to select the fraction of the cycle during which plate current flows, and to calculate grid bias and exciting voltage. The fraction of the cycle is determined as a compromise between a number of conflicting factors, since a small value gives high plate efficiency but results in small power output and large driving power, while a large value gives a large output, but makes the plate efficiency low. The factors normally balance when the current flows for something between 90 and 180 electrical degrees.

The grid bias E_c required to cause the plate current to flow for θ electrical degrees is given by the equation[8]

$$\text{grid bias} = E_c = \frac{E_b}{\mu} + \left(E_{\max} + \frac{E_{\min}}{\mu}\right)\frac{\cos\theta/2}{1-\cos\theta/2} \tag{2}$$

where $E_{\max}$, $E_{\min}$, E_b, and μ have the same definitions as above. With the grid bias E_c and the maximum positive grid potential $E_{\max}$ both known, the exciting voltage is $(E_{\max}+E_c)$.

Fourth. One is now ready to calculate the power relations. The first step is the determination of the components of the total space current for the I_m and the angle of flow selected above. This is done with the aid of the curves in Fig. 2 using a value of the exponent α which experience indicates is desirable, and which in the absence of information to the contrary can be assumed to be three halves.

The total space current determined in this way is divided between the plate and grid electrodes of the tube. In order to determine the effect of the grid current it is necessary to make an estimate based on experience, of the fraction of the total direct space current that will be diverted to the grid. This percentage is commonly ten to twenty-five per cent in air-cooled tubes operating at plate potentials up to several thousand volts, while with water-cooled tubes it is less, with the grid

[8] This equation follows from the fact that at the instant the plate current stops flowing the signal and load voltages are $\theta/2$ degrees from their crest values, so that at this instant the effective anode voltage is

$$\left(\frac{E_b-(E_b-E_{\min})\cos\theta/2}{\mu} - E_c + (E_{\max}+E_c)\cos\theta/2\right),$$

and this must equal zero.

current often reversing as a result of secondary electron emission. After the probable grid current has been estimated, the direct plate current is obtained by subtracting the direct grid current from the direct-current component of the total space current I_m. In the event that the grid current is negative, the direct component of the plate current will be larger than the direct-current component of the total space current.

The fundamental alternating-current component of the total space current is likewise divided between grid and plate electrodes, with the amount going to the grid very nearly equal to twice the direct current component of the grid current. This comes about because most of the grid current flows during the very crest of the cycle, and as shown in a previous paper[7] this is equivalent to an alternating component that is twice the direct-current value of the grid current. The alternating component of the plate current is hence the alternating-current component of the total space current as obtained from Fig. 2 minus twice the direct grid current. The power input to the class C amplifier is now the product of battery voltage and direct plate current, or

$$\text{power input} = E_b \times I_{dc} \tag{3}$$

where I_{dc} is the direct plate current. Likewise the power delivered to the load is equal to half the product of alternating plate current and alternating voltage developed across the load, or

$$\text{power output} = \frac{(E_b - E_{\min})I_{ac}}{2} \tag{4}$$

where I_{ac} is the crest value of the fundamental frequency component of the plate current. The plate dissipation is the difference between these two powers and the efficiency is their ratio.

The grid driving power is then approximately equal to the direct grid current as estimated above times the crest value of the exciting voltage.[9]

Fifth. The above steps give a complete solution of the class C amplifier for the assumed values of I_m, θ, $E_{\max}$, and $E_{\min}$. If the results obtained are not satisfactory, one can make a new choice of initial operating conditions and recalculate to obtain more nearly the optimum performance.

[9] This is because to a first approximation the grid current can be assumed to flow when the grid exciting voltage is at its crest. *See* F. E. Terman, "Radio Engineering," p. 234. Experimental work indicates that power calculated in this way is in the order of five to ten per cent high. *See* H. P. Thomas, "Determination of grid driving power in radio-frequency power amplifiers," Proc. I.R.E., vol. 21, pp. 1134–1141; August, (1933).

Example

In order to show in detail how the above procedure is carried out, a class C amplifier will now be designed using a type 800 tube at 1000 volts plate potential. The complete characteristic curves of such a tube as given by the manufacturer are shown in Fig. 3.

The peak emission I_m will be taken as 407 milliamperes. This is arrived at by assuming that initially the thoriated tungsten filament is capable of emitting 100 milliamperes per watt of heating power, and then allowing a factor of safety of six to provide for deterioration dur-

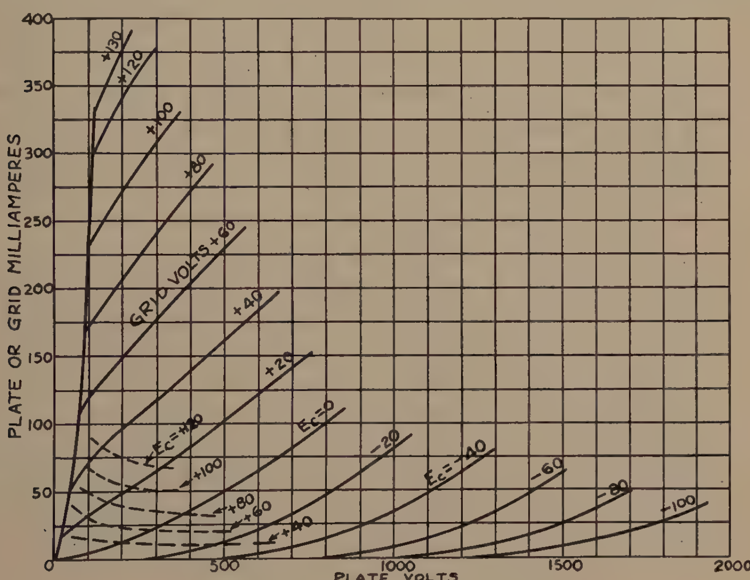

Fig. 3—Characteristic curves of type 800 tube.

ing life. With a small tube such as the 800, $E_{\min}$ and $E_{\max}$ will normally be about equal. Assuming this equality and referring to Fig. 3, it is found that 123 volts on both grid and plate will draw a space current of 407 miliamperes. The next step is the selection of the angle of current flow. A value of 120 degrees represents a reasonable compromise between high efficiency and large output and will be tentatively selected. From the curves of Fig. 2, on the assumption that $\alpha = 3/2$, the factors for direct plate current and alternating plate current are 0.19 and 0.35, respectively, so that the corresponding components of the total space current are $407 \times 0.19 = 77.5$ milliamperes direct current, and $407 \times 0.35 = 142$ crest milliamperes of fundamental frequency. It is now necessary to make allowance for the part of the total space cur-

rent diverted to the grid. Assuming that twenty per cent of the total direct space current will be diverted to the grid as reasonable for small tubes, the direct grid current will be 15.5 milliamperes. The direct plate current is then $77.5-15.2=62$ milliamperes, and the fundamental frequency component of the plate current is similarly $142-2\times15.5=111$ milliamperes crest value.

The power input to the plate circuit is the product of direct plate current and plate voltage, or $1000\times0.062=62$ watts, while the power output is half the product of crest alternating plate current and crest alternating voltage across the load, and so is $0.111\ (1000-123)/2=49$ watts. The plate loss is $62-49=13$ watts, and the efficiency is $49/62=79$ per cent.

The grid bias required as calculated by (2) is found to be 198 volts. The crest alternating driving voltage is $(E_c+E_{\max})$ or 321 volts, and the grid driving power is to a first approximation $321\times0.0155=5.0$ watts. The load impedance that is required is the ratio of alternating voltage $(E_b-E_{\min})$ to the alternating-current component of the plate current, and so is $(1000-123)/0.111=7900$ ohms.

If the above results do not represent the desired operating conditions, one can readily make a new set of calculations on the basis of a new value of angle of current flow, or a different combination of $E_{\min}$ and $E_{\max}$, or both. In particular it will be noted that although the above operating conditions develop the normal rated output, the plate losses are considerably lower than the maximum allowable loss, and that increased output at a somewhat lower efficiency could be obtained by using a larger angle of current flow. Also, by making $E_{\min}>E_{\max}$, one could reduce the grid current, and hence the driving power.

Accuracy of Calculations

The only approximations involved in the method of analyzing class C amplifiers outlined above are the uncertainty regarding the exact amount of grid current, and the assumption that the exponent α in (1) is constant. The necessity of making a guess as to the grid current need not introduce appreciable error since the grid current is always a small proportion of the total space current, and therefore a considerable percentage error in grid current represents only a very small percentage error in the plate current components. Also any error in estimating grid current alters the calculated power input, power output, and tube losses, all in substantially the same proportions and therefore has little effect on the plate efficiency.

The assumption that the exponent α in (1) is constant is found to be substantially correct over the essential part of the tube characteris-

tics provided the peak space current I_m is not so great as to approach saturation by insufficient electron emission. In the case of tungsten filament tubes operated with total space currents very close to the peak emission available, some saturation effects are normally found. Even then the error that results is not particularly great, and alters both power input and power output in about the same proportion, so that the predicted plate efficiency will still be almost exactly correct.

With a little experience it is possible to make fairly accurate allowances for these incipient saturation effects. Thus when the value of I_m is taken from complete characteristic curves any tendency toward saturation that is present will cause the calculated input and output powers to both be slightly low. On the other hand, when one extrapolates a curve of I_m as a function of $(E_g+E_{p/\mu})$ to get the operating conditions required to draw the maximum space current I_m, the calculated input and output powers will both be slightly higher than the true values.

The accuracy obtainable with analyses based upon the curves of Fig. 2 is shown by the following examples.

Example One. A point-by-point calculation of the performance of the class C amplifier considered above, using the same $E_{\min}$, $E_{\max}$, and angle of current flow, gives the following results:

	By analysis using Fig. 2	By exact point-by-point calculation
Direct plate current	62. ma	63.9 ma
Direct grid current	15.5 ma	12.6 ma
Alternating plate current	111. ma	116. ma
Power input	62 watts	64 watts
Power output	49 watts	51 watts
Plate loss	13 watts	13 watts
Plate efficiency	79 per cent	79.5 per cent

It will be noted that if the grid current had been more accurately estimated the agreement would have been practically perfect.

Example Two. Mouromtseff and Kozanowski[2] have given on page 761 of their recent paper the results obtained from a point-by-point analysis of a tube having characteristics such as shown in their Fig. 1, under the following conditions:

$$E_b = 20{,}000 \text{ volts}$$
$$E_{\min} = 6{,}500 \text{ volts}$$
$$E_{\max} = 740 \text{ volts}$$
$$\text{Grid bias} = 1{,}400 \text{ volts}$$
$$\mu = 20$$
$$\text{Direct grid current} = -100 \text{ milliamperes}$$

From these data, the angle of current flow is found by (2) to be 148 degrees. Reference to the characteristic curves given for the tube shows that the maximum space current is 7.2 amperes. With these data, the results calculated with the aid of the curves of Fig. 2 for $\alpha=3/2$ are given in the following tabulation, together with the exact results calculated by Mouromtseff and Kozanowski:

	By Fig. 2	Exact	Percentage Difference
Input power	35.5 kw	37.1 kw	−4.3 per cent
Output power	21.0 kw	21.75 kw	−3.4 per cent
Plate loss	14.5 kw	15.35 kw	−5.5 per cent
Efficiency	59.1 per cent	58.8 per cent	+0.5 per cent

It is apparent from the above examples that the degree of accuracy obtainable is quite satisfactory for preliminary calculations and can be safely used as the basis for circuit design.

Practical Adjustment of Class C Amplifiers to Realize the Desired Conditions

After the desired conditions have been calculated, one still must realize these in actual operation. The first step is to obtain the appropriate grid bias and plate supply potentials, after which the load is coupled into the plate circuit a reasonable amount. If a positive peak voltmeter is available the excitation is adjusted to give the desired positive grid potential E_{max}, after which the load coupling is varied until a point is found where the direct plate current approximates the calculated value, and further reduction in coupling causes the grid current to increase rapidly and become excessive. If a positive peak voltmeter is not employed, the proper procedure is to adjust the grid excitation until the total space current I_p+I_g approximates the desired value, after which the load coupling is adjusted as above. It may then be necessary to readjust the excitation and load coupling slightly to realize the desired total space current and direct plate current.

The amount the grid is driven positive can be measured directly by means of a peak vacuum tube voltmeter such as illustrated in Fig. 4. This instrument has been described elsewhere[7] so will be considered here only briefly. It makes use of a diode tube, preferably an 879 or similar tube capable of standing a high voltage and having relatively low interelectrode capacity. The anode is connected directly to the grid of the class C amplifier while the cathode is biased positive with respect to the amplifier cathode by the potentiometer P until the milliammeter M just begins to show current. Under this condition the

cathode potential is substantially equal to the most positive potential reached by the rectifier anode, so that voltmeter V then reads E_{max} directly.

If desired it is also possible to arrange a modified peak voltmeter (trough meter) which is capable of reading the minimum plate potential E_{min}, but this is not really required. Voltmeters for this purpose are described elsewhere.[10]

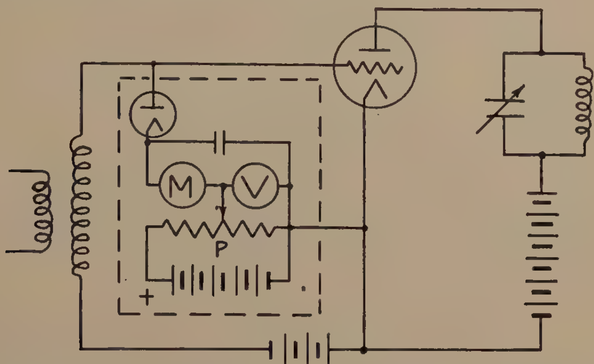

Fig. 4—Peak voltmeter for adjusting E_{max} to desired value.

Appendix

The curves shown in Fig. 2 can be calculated by the following procedure:

The waves to be analyzed have a shape corresponding to (1). When the exponent α is unity this is essentially a section of a sine wave having a duration θ and a height above the base line of $I = I_m$, as shown in Fig. 5. The equation of such a wave is

$$i = I \frac{(\cos\beta - \cos\theta/2)}{1 - \cos(\theta/2)} \quad \text{for } \beta < \theta/2 \tag{5}$$

$$i = 0 \quad \text{for } \beta > \theta/2$$

where i is the amplitude above the axis at β degrees from the crest. If the exponent α in (2) is not unity, then the equation of current is simply the right-hand side of (5) raised to the α power, with I_m substituted for I^α. That is,

$$i = I_m \left[\frac{\cos\beta - \cos\theta/2}{1 - \cos\theta/2}\right]^\alpha \quad \text{for } \beta < \theta/2 \tag{6}$$

$$i = 0 \quad \text{for } \beta > \theta/2.$$

[10] F. E. Terman, "Measurements in Radio Engineering," McGraw-Hill Book Co.

The value I_{dc}/I_m is then the right-hand side of (6) averaged over a cycle, and divided by I_m, or

$$\frac{I_{dc}}{I_m} = \frac{1}{\pi}\int_0^{\beta=\theta/2}\left[\frac{\cos\beta - \cos\theta/2}{1 - \cos\theta/2}\right]^\alpha d\beta. \tag{7}$$

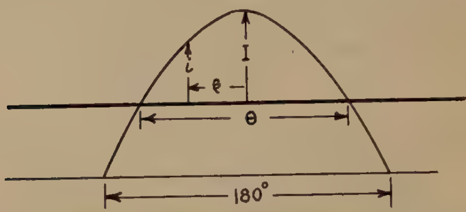

Fig. 5—Figure illustrating notation in equation (5).

The fundamental frequency component of the wave of (6) is found in the usual Fourier method by multiplying by cos β and averaging over the cycle. This yields

$$\frac{I_1}{I_m} = \frac{2}{\pi}\int_0^{\beta=\theta/2}\left[\frac{\cos\beta - \cos\theta/2}{1 - \cos\theta/2}\right]^\alpha \cos\beta\, d\beta. \tag{8}$$

The integrations involved in (7) and (8) cannot be carried out in a simple mathematical manner for values of α that are not integers. The authors therefore used point-by-point methods in deriving Fig. 2 and after trying out several procedures believe that Simpson's rule is the most satisfactory. According to this, the base θ of the curve is divided into n equal parts, where n is an even number. The area under the curve is then

$$\text{area} = \tfrac{1}{3}h\left[y_0 + 4(y_1 + y_3 + y_5 \cdots y_n - 1) + 2(y_2 + y_4 + y_6 \cdots y_n - 2) + y_n\right] \tag{9}$$

where h is the distance between adjacent ordinates ($h=\theta/n$), and y_0, y_1, y_2, etc., are the heights of the curve for the various positions along the base of the curve. The accuracy increases as the number of intervals n increases. Enough points were used in obtaining Fig. 2 to insure an error of less than one per cent.

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DESCRIPTION AND CHARACTERISTICS OF THE END-PLATE MAGNETRON*

By

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Summary—*A new type of magnetron is described which is especially adapted to the generation of centimeter waves. It possesses several advantages over the simple magnetron, namely: (1) greater stability with respect to fluctuations of supply voltages, (2) less tendency to oscillate at undesired long wavelengths, (3) greater efficiency, (4) greater output, and (5) greater ease of adjustment. Static and dynamic characteristics are discussed. The effect of space charge on electron motion and tube performance is treated mathematically, and supporting experimental data are presented. Evidence is given that for best operation an optimum space-charge condition is required, which can conveniently be established and maintained by the use of end plates. Power output is limited by a type of instability involving electron bombardment of the filament, and apparently initiated by excessive space charge.*

Introduction

In THE January 1935, number of the Proceedings, an article was published[1] on the transmission and reception of centimeter waves, which gave a general description of apparatus, technique, and results, but went into no great detail. The present article deals specifically with the magnetron oscillator, which was used, and some of the problems associated with electronic oscillators in general.

Description of Tube

The end-plate magnetron has proved comparatively reliable and efficient with regard to other types of oscillators in the centimeter region. From a single tube, 2.5 watts at 3000 megacycles has been obtained with an efficiency of twelve per cent with respect to plate dissipation. With suitably regulated power supply it has been operated for days at a time without requiring readjustment.

A photograph of the tube is shown in Fig. 1, and a simplified sketch of the elements in Fig. 2. The plate consists of a tantalum or molybdenum cylinder, split longitudinally into two equal halves. Its length is about eight millimeters, diameter four millimeters, and slit width 0.5 millimeter. A pure tungsten filament of diameter 0.127 millimeter is positioned along the axis of the cylinder. At each end of the cylinder, from 0.5 to 1.0 millimeter distant, is an end plate in the form of a

* Decimal classification: R331. Original manuscript received by the Institute, July 1, 1935.

[1] Wolff, Linder, and Braden, Proc. I.R.E., vol. 23; pp. 11–23, January, (1935).

flat disk about six millimeters in diameter and having a half-millimeter hole at the center through which the filament passes. Only one end plate is shown in the sketch in order not to obstruct the view of the other elements. To each of the cylinder halves is connected one side of

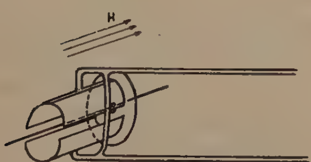
Fig. 1—End-plate magnetron.

the transmission line. A shorting bar is welded across the line at a distance of about 0.5 millimeter from the nearest edge of the cylinder halves.

Accuracy of alignment of the elements is of the greatest importance. The straightness and positioning of the filament along the central axis is especially critical, and departures of a few thousandths of an inch

greatly reduce the output. The end plates should be as close as possible to the ends of the plate without danger of a short circuit. All internal parts should be nonmagnetic to avoid distortion of the magnetic field. However, the use of nickel wires to support the elements seems not very detrimental. A high vacuum is essential to stable operation. The use of a getter is desirable, but care should be taken to confine it to parts of the tube remote from the oscillating elements.

The circuit employed is shown in Fig. 3. It consists of a high voltage source S with a voltage divider R for supplying plate and end-plate voltages. Modulation is most simply obtained by means of a transformer M in series with the plate. Oscillations can be produced with voltages of from a few hundred to about 1500 volts applied to the plates. The plate lead is conveniently connected to the center of the



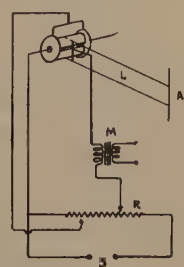


Fig. 2—Elements of end-plate magnetron.

Fig. 3—Connections for operation with plate modulation.

shorting bar, which is a voltage node. The optimum potential for the end plates varies somewhat according to the spacing between them and the cylindrical plates, but is usually from one half to three quarters that of the plates. For continuous, stable operation a filament emission of about three milliamperes is suitable. For maximum output an emission of about eight to ten milliamperes is required. A magnetic field must be applied parallel to the axis of symmetry of the elements. For a frequency of 3000 megacycles, this should have a strength of about 1400 gauss.

Static Characteristics

The static characteristics and electron motion for the case of a simple magnetron consisting of a cylindrical plate with concentric filamentary cathode, have been studied by Hull[2] both experimentally and mathematically, for the case in which the magnetic field is codirectional with the axis of symmetry of the elements. In the absence

[2] A. W. Hull, *Phys. Rev.*, vol. 18, p. 31, (1921).

of a magnetic field, electrons travel along straight lines radially from the cathode to the plate as in Fig. 4 (a). When a weak magnetic field is applied the paths become curvilinear, Fig. 4 (b). At some stronger value of the field, called the critical cutoff value, the path curvature

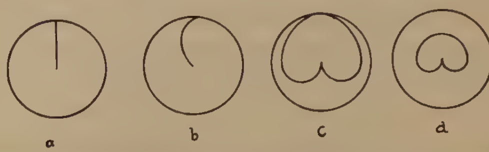


Fig. 4—Effect on electron trajectories of increasing magnetic field.

becomes so great that the electrons no longer reach the plate but graze it tangentially, curve around and return to the cathode, as in Fig. 4 (c). Further increases in field strength cause still greater curvature so that the dimensions of the path become smaller, as in Fig. 4 (d).

Because of this action of the field, the plate-current characteristic is as shown in Fig. 5, the current dropping to zero at the critical value H_c. The sharpness of this cutoff depends, among other things, upon the accuracy of alignment of the tube elements, the velocity of emission of the electrons, and the amount of space charge. In actual cases the cutoff is never as sharp as shown, especially when the dimensions of the elements are small, and the voltages large. Fig. 5 is applicable to either the simple magnetron or the end-plate type.

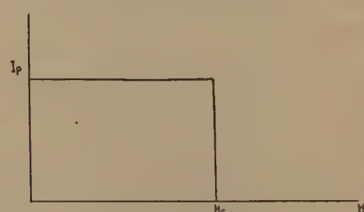


Fig. 5—Idealized curve for plate current vs. magnetic field.

The addition of end plates at a positive potential causes an important change in the electron motion. A velocity component parallel to the cathode is added, causing the electron to traverse a helical path such that it is pulled away from its point of emission towards the end plates. Due to the shielding effect of the cylindrical plate, the field of the end plates does not extend very far into the cathode-anode space, and its strength is small in most of that region in comparison to that of the regular plate. For this reason the electron motion is not radically affected unless the magnetic field exceeds the critical value. Above this critical value the trajectory becomes a long helix extending

from the point of emission to the end plates. The current characteristic is shown in Fig. 6. As H passes through the critical value H_c, the plate current cuts off, in the same manner as in Fig. 5, simultaneously the end-plate current increases. The end-plate current is usually not as great as the plate current, probably because of space-charge limitation of current from the more remote parts of the filament.

So far only cases have been considered in which the magnetic field is codirectional with the axis of symmetry of the tube elements. Cases in which this is not so are important since it has been found that a tilt of a few degrees is usually necessary for the simple magnetron in order to obtain maximum output.

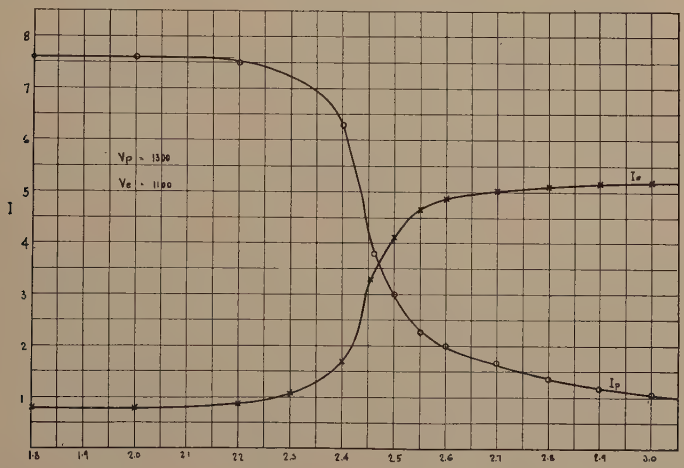


Fig. 6—Experimental curves for plate and end-plate currents vs. magnetic field.

If the magnetic field makes an angle with the axis, the electron trajectories will no longer be plane curves as in Fig. 4, but will be helices, the electrons traveling outwards towards the ends of the cylindrical plate. Thus, inclining the tube has an effect on the electron motion similar to that of end plates at a positive potential. The path of an electron in such a tilted tube is quite complicated, and has not been worked out mathematically for the cylindrical plate case. However, qualitative considerations and comparison with the parallel plate case indicate that the axes of these helices lie parallel to the direction of the magnetic field.

The parallel plate case is useful, and easily solved. Chose coordinates as in Fig. 7, x_1 coinciding in direction with the magnetic field H. The differential equations are

$$\left.\begin{aligned} m\ddot{x}_1 &= Ee \sin\theta, \\ m\ddot{x}_2 &= Ee\cos\theta - He\dot{x}_3, \\ m\ddot{x}_3 &= He\dot{x}_2 \end{aligned}\right\} \quad (1)$$

and,

The solutions are

$$\left.\begin{aligned} x_1 &= R\omega^2 \sin\theta\cdot\frac{t^2}{2}, \\ x_2 &= R\cos\theta(1-\cos\omega t), \\ x_3 &= R\cos\theta(\omega t - \sin\omega t), \end{aligned}\right\} \quad (2)$$

and,

where,

$$R = \frac{Em}{H^2e} = \text{the radius of the generating} \quad (3)$$

circle of the cycloid, and

$$\omega = \frac{He}{m}. \quad (4)$$

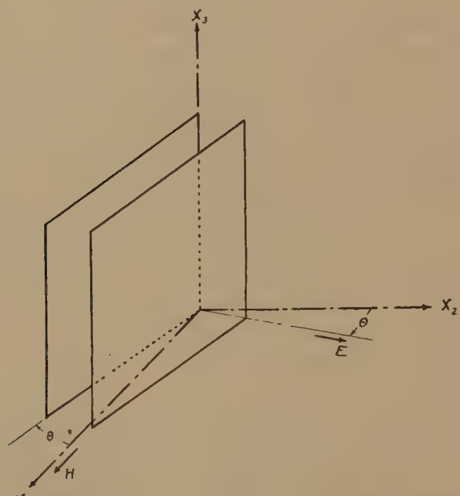


Fig. 7—Plates of flat-plate magnetron with co-ordinate system. Note that the plates have been rotated about the x_3 axis through an angle θ in the x_1 x_2 plane.

This solution represents a cycloidal helix whose axis lies along x_1; i.e., in the direction of H. Tilting of a magnetron at an angle θ with respect to H, therefore, causes the electrons to travel along helices whose axes make an angle θ with the filament. Hence, if before tilting no current reached the plate because of cutoff due to the magnetic field, sufficient tilting will cause a current to flow.

The variation of the plate current I_p as the angle θ is changed, is shown in Fig. 8. This curve is valid for either the simple or end-plate

magnetron. In the latter case there is a simultaneous variation of end-plate current I_e, which also is given in the figure.

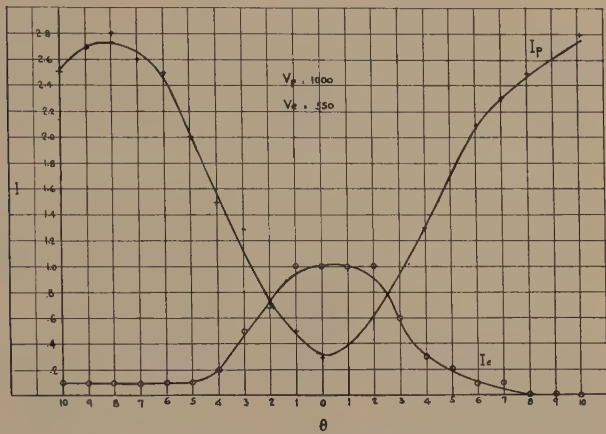
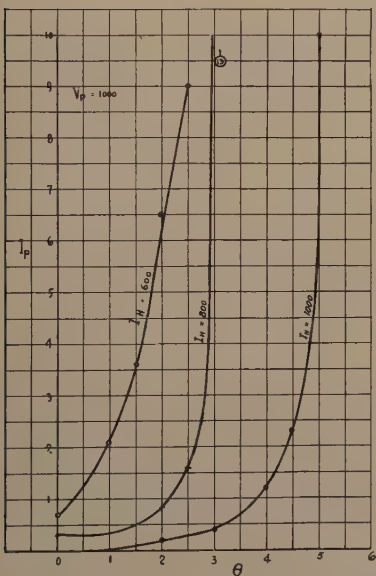


Fig. 8—Variation of plate and end-plate currents as magnetron is tilted in magnetic field.



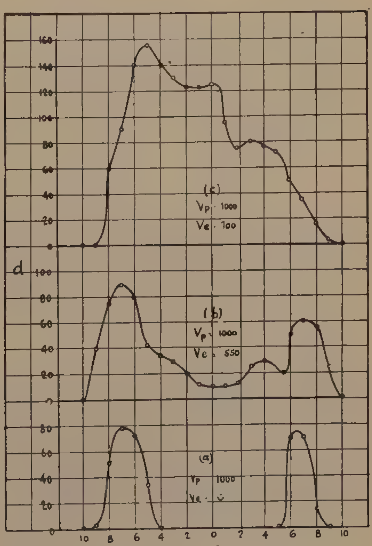


Fig. 9—Variation of plate current with angle of tilt for several values of magnetic field.

Fig. 10—Elimination of necessity of tilting tube by increasing end-plate potential. Output power is plotted against angle of tilt for various end-plate voltages.

The effect of varying the magnetic field strength H, on the shape of this curve is shown in Fig. 9. It is evident that for stronger magnetic

fields a greater tilt is required to obtain a given plate current. This is due to the decrease in the dimensions of the electron trajectory caused by increasing H, the change in dimensions being such that the electron does not approach so closely to the plate.

The relation between V, H and the maximum radial distance r_1 from the cathode may be derived from the expression given by Hull,[2]

$$\left(\frac{dr}{dt}\right)^2 = \frac{2eV(r)}{m} - \left(\frac{He}{2m}\right)^2 r^2. \tag{5}$$

The maximum value of r is obtained by equating this to zero. This yields,

$$r_1 = \left(\frac{8mV(r_1)}{H^2e}\right)^{1/2}. \tag{6}$$

The function $V(r)$ is arbitrary. For cases in which we are interested the potential distribution may be represented sufficiently well by the expression,

$$V = V_p\left(\frac{r}{r_p}\right)^n, \tag{7}$$

where V_p is the plate potential, and r_p its radius. The exponent n is usually fractional, being equal to two thirds for the space-charge-limited case. Substituting (7) in (6) yields

$$\frac{r_1}{r_p} = \left(\frac{8mV_p}{H^2er_p{}^2}\right)^{1/(2-n)}, \tag{8}$$

which is the desired relation. Evidently r_1 decreases as H increases in agreement with the experimental results.

Dynamic Characteristics

Theories of the operation of the magnetron oscillator will not be discussed at length here. Several good papers have appeared[3,4,5] but there is as yet no general agreement as to the mechanism of oscillation. Apparently there are several modes of oscillation. That with which this article is concerned involves an electron transit time from cathode to plate of substantially one half a period. Due to this delayed arrival at the plate, the alternating-current component of electron current is out of phase with the voltage, giving rise, under certain conditions, to negative resistance. The phase angle which gives maximum negative

[3] H. G. Moeller, *E.N.T.*, vol. 7, pp. 293 and 411, (1930).
[4] W. E. Benham, *Phil. Mag.*, vol. 11, p. 457 (1931); *Proc. Phys. Soc.*, vol. 47, p. 1, (1935).
[5] E. C. S. Megaw, *Jour. I.E.E.*, (*London*) vol. 72, p. 326, (1933).

conductance appears to depend upon the amount of space charge. For zero space charge the angle is 180 degrees; for space-charge limitation it is near 270 degrees.[4]

On account of the cyclic motion of the electrons, their closest approach to the plate occurs at times $(2m-1)T_1$, where $m=1, 2, 3, \cdots$, and T_1 is the time required for an electron to travel from the cathode to its point of first close approach to the plate. Maximum output occurs if the electrons just graze the interior surface of the plate, as in Fig. 4(c). These two conditions can be met by suitably adjusting the magnetic field and plate voltage, the former governing chiefly the frequency of electron revolution, and the latter its maximum radius of excursion from the cathode.

As has been mentioned, a tilt of several degrees, or the use of positive end plates is usually required for the greatest efficiency. As the end-plate potential is increased, the output at $\theta=0$ increases, until eventually the maximum output can thus be obtained without tilting. The manner in which the end plates thus eliminate tilt is shown in Fig. 10. In curve (a) the end-plate potential is zero and the tube operates essentially as a simple magnetron. Maximum output is obtained for $\theta=\pm 7$ degrees. There is no output at $\theta=0$. If the end-plate potential is raised to 550 volts, curve (b) is obtained. Here is is observed that the two peaks are broader and there is some output at $\theta=0$. Finally if $V_e=700$, the two side peaks disappear, and there is substantially only one broad peak, whose center would presumably be at $\theta=0$, if the tube were perfectly symmetrical.

In Fig. 11 is shown the variation of radiated energy with V_e, all other factors being held constant. This curve is not always as simple and symmetrical as shown. Subsidiary peaks appear if the various oscillating elements, including the electron cloud, are not properly adjusted to each other.

In the above-described tests both end plates have been at the same potential. It has been found that sometimes slightly better results are obtained if they are operated at different potentials. This is probably due to dissymmetry. In general, however, such differences in potential have no great effect, so that a tube with only one end plate is almost as good as one with two.

Several forms of end plates have been tried. These included rings of wire, and cylinders of the same size as the plate but shorter. The disk form was found best.

No satisfactory explanation has heretofore been offered for the necessity of tilting. Megaw[5] once suggested that it was necessary in order to compensate for the potential drop along the filament, but later

experiments with an equipotential cathode caused him to abandon this hypothesis. Dissymmetry in the form of the elements was suggested as an explanation by Slutzkin and Steinberg,[6] but the fact that tilts in either direction from zero are practically equally effective, as well as the results obtained with end plates and no tilt, contradict this. Kilgore[7] has made the more plausible suggestion that tilting reduces space charge since it enables electrons eventually to reach the plate instead of cycling around indefinitely. The employment of positive end plates would have a similar reducing effect on space charge because of the current which they collect.

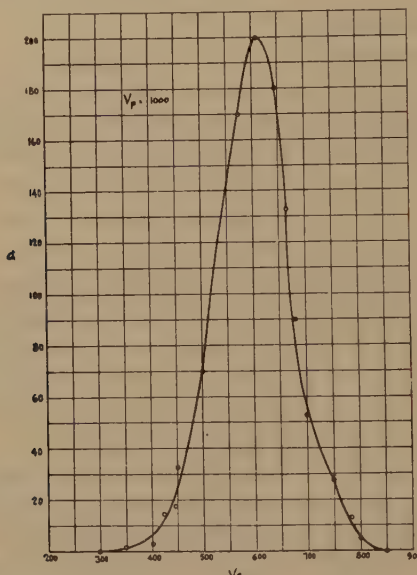


Fig. 11—Variation of power output with end-plate potential.

That the true function of the end plates, or tilting, is thus to reduce space charge is well supported by results obtained by the writer. It appears that for best operation it is always necessary to provide a means whereby a certain portion of the electrons may be drawn out of the cathode-plate region, thus maintaining an optimum space-charge condition.

The writer has succeeded in accomplishing this withdrawal of electrons in three different ways: (1) by tilting, (2) by positive end plates, and (3) by high plate voltages. These will be discussed in turn.

[6] Slutzkin and Steinberg, *Ann. der Physik*, vol 1, p. 658, (1929).
[7] G. R. Kilgore, Proc. I.R.E., vol. 20, pp. 1741–1751; November, (1932).

The manner in which electrons are enabled to reach the plate when the tube is tilted has been described in the section on static characteristics. It remains to discuss how the tilt affects output, especially in connection with the maintenance of the optimum space-charge condition.

It has been shown that the plate current I_p is a function of θ rising from practically zero at $\theta=0$ to a maximum on both sides (see Fig. 8). Hence at some definite value of θ, either positive or negative, the requisite number of electrons can usually be withdrawn to establish the optimum space-charge state and yield maximum output. This accounts for the two peaks shown in Fig. 10 (a) and (b).

A further test of this theory may be made by studying the effect on the optimum angle of varying the plate voltage. It is evident from (8) that an increase in V_p results in an increase in r_1, so that in consequence the electrons approach closer to the plate. Because of this closer approach, a smaller tilt is required to enable a given proportion of them to reach the plate. Hence the optimum space-charge condition occurs at smaller angles for larger plate voltages. The curve obtained experimentally is shown in Fig. 12, in which θ_{opt} is seen to decrease continuously as V_p increases, in agreement with the above theory. The power output is indicated by the curve (d).

In the case where end plates are used, and the tube is not tilted, it is found experimentally that maximum output is obtained over most of the voltage range, for a substantially constant ratio of end-plate voltage to plate voltage. This ratio, indicated by R_1, is plotted against V_p in the lower part of Fig. 13. It appears that this constant ratio is necessary to yield the required space-charge condition. The output corresponding to R_1 is given by d_1 in the upper section of the figure.

If the plate voltage V_p is increased beyond a certain point (about 970 volts in the present case) the optimum ratio of V_e/V_p changes radically. This leads to the third of the above-mentioned methods for removing electrons from the interelectrode space; i.e., high plate voltages. The optimum ratio for this voltage region is given by R_2 and the corresponding output by d_2. Obviously much lower values of end-plate voltage are required than previously, in fact for $V_p=1150$, the optimum end-plate voltage is zero. This is explained as due to the collection of electrons by the plate itself. For high values of V_p the electrons graze the plate so closely that appreciable numbers of them are eliminated by the plate itself. Hence as V_p increases, elimination by the end plates becomes increasingly unnecessary, and eventually undesirable. The increase in plate current due to the closely grazing electrons is shown by the curve I_p in the lower part of Fig. 13. On the

section of the curve where double points are shown, the ones surrounded by squares correspond to R_2. The output corresponding to R_2 is shown by d_2 in the upper section.

The state of operation corresponding to the peak of the d_2 curve for which $R_2=0$, is the same as that shown in Fig. 12 at $\theta=0$. Hence by sufficiently increasing V_p, both the simple magnetron and the end-plate magnetron can be brought into the same oscillating condition;

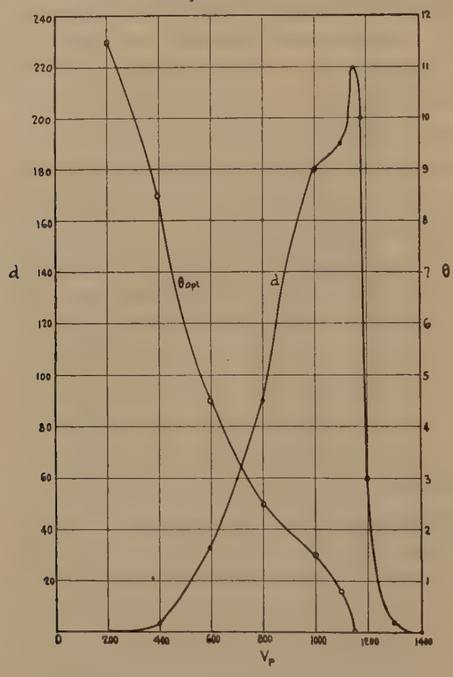


Fig. 12—Variation of power output and optimum angle of tilt with plate voltage.

that is, a state in which the plate itself eliminates a sufficient number of electrons to produce the optimum space-charge condition. For lower plate voltages, electron elimination must be effected by other means such as end plates or tilting.

A fact of considerable practical importance arises from the constancy of the ratio R_1 over most of the voltage range, as shown in Fig. 13. This gives the end-plate magnetron unusual constancy of output, with respect to accidental or other variations of plate supply voltages, since a constant ratio is easily maintained in spite of such variations. In order to attain the same result with a simple magnetron, it would be necessary to vary the angle of tilt automatically so as always to have the optimum value at all plate voltages.

A comparison of the two types of tubes in this respect is given in Fig. 14. Curves (a) and (b) are for the end-plate tube for constant V_e/V_p ratios as indicated. Curves (c) and (d) are for the same tube operated as a simple magnetron; i.e., $V_e=0$ and $\theta\neq 0$. It will be noted that the end-plate tube oscillates over a wider voltage range. The voltage range for the tilted tube is especially narrow at low plate voltages, as shown by curve (c).

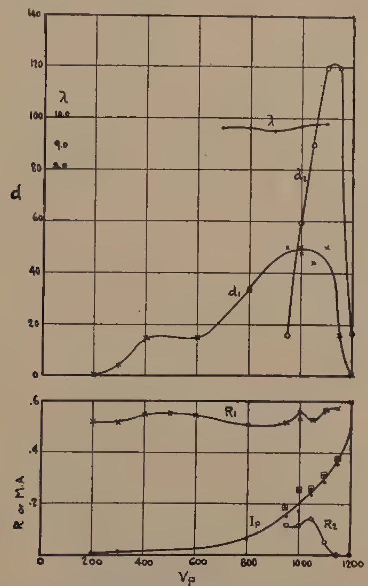


Fig. 13—Curves showing approximately constant ratio (R_1) of end-plate to plate voltage over most of operating voltage range; also corresponding values of wavelength, power output, and plate current.

Effect of Space Charge on Electron Motion

The results of the preceding section indicate quite clearly that the function of end plates, or tilting, is to remove a certain proportion of electrons from the interelectrode space and thus create an optimum space-charge condition. However no suggestion has yet been offered, and it is not clear, as to why a definite space-charge condition should be desirable.

Space-charge alters the form of the potential distribution so that the electric field strength near the filament is reduced. This results in a change in the shape of the electron trajectory, such that in general its velocity is less, its curvature greater in this region, and its path longer. In certain cases these changes may be very pronounced, in fact, as will be shown below, the motion may become aperiodic.

Space-charge effects can be conveniently discussed by making use again of the expression for potential

$$V = V_p\left(\frac{r}{r_p}\right)^n, \tag{7}$$

in which n may be regarded as a parameter determining the degree of

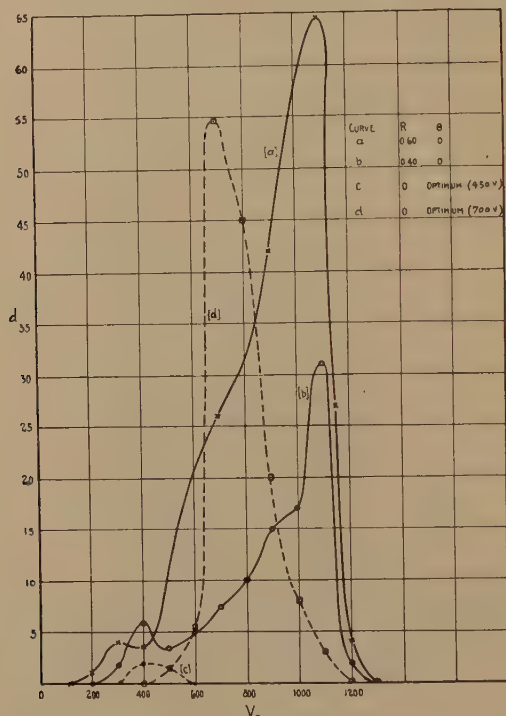


Fig. 14—Comparison of simple magnetron and end-plate magnetron. Power output vs. plate voltage.

space charge. Of course not all potential distributions which actually occur can accurately be represented by (7). However, most of them can be thus approximated with sufficient accuracy, and near the space-charge-limited condition the approximation is especially good.

In Fig. 15, equation (7) is plotted for values of n ranging from $+\infty$ to $-\infty$. The vertical scale is linear in the lower half of the figure, but in the upper half a reciprocal scale is used. The upper section is not of particular interest to the present problem, but is included for the sake of completeness.

Negative values of n yield potential distributions which are infinite at the origin. These can be approximated in practice by potentials which jump from zero to a high value greater than V_p near the filament, and then drop off, down to V_p at $r=r_p$. Such distributions could be obtained with a grid at high positive potential near the cathode.

When n is zero, the potential has the constant value V_p throughout the cathode-plate region. If n has values between zero and unity the field is very strong near the filament and drops off to low values at the

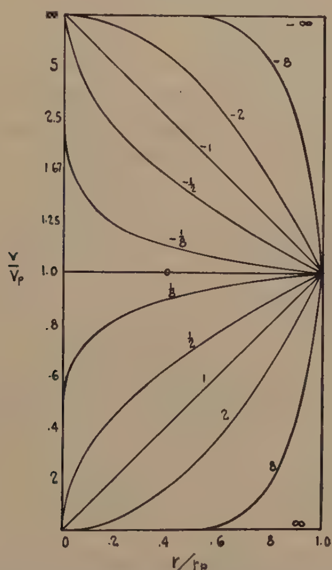


Fig. 15—Theoretical potential distributions given by $V=V_p(r/r_p)^n$ for values of n from $-\infty$ to $+\infty$.

plate. If n is unity the field is constant everywhere. For n between unity and infinity the field is zero at the cathode and high at the plate.

Introducing (7) in the general equation for radial motion (5), yields,

$$\left(\frac{dr}{dt}\right)^2=\frac{2eV_p}{m}\left(\frac{r}{r_p}\right)^n-\left(\frac{He}{2m}\right)^2 r^2.$$

This can be written in terms of r/r_p, thus,

$$\frac{d}{dt}\left(\frac{r}{r_p}\right)=\sqrt{\frac{2eV_p}{mr_p{}^2}\left(\frac{r}{r_p}\right)^n-\left(\frac{He}{2m}\right)^2\left(\frac{r}{r_p}\right)^2}. \tag{9}$$

The solution is

$$\frac{r}{r_p} = \left[a \sec \left(\frac{a(n-2)(bt+C)}{2} \right) \right]^{2/(n-2)}, \tag{10}$$

where,

$$a = \left(\frac{H^2 e r_p^2}{8mV_p} \right)^{1/2},$$

$$b = \left(\frac{2eV_p}{mr_p^2} \right)^{1/2},$$

and C is an integration constant.

Now if $t=0$, $r=0$. This is satisfied by (10) only if $n<2$, since the secant is never zero. However, if $n<2$ the exponent is negative, and the secant may be replaced by its reciprocal, the cosine.

The integration constant C is found to be $\pi/a(n-2)$. Hence the solution is

$$r = r_p \left[\left(\frac{8mV_p}{H^2 e r_p^2} \right)^{1/2} \sin \frac{(2-n)He}{4m} t \right]^{2/(2-n)}. \tag{11}$$

Consider first the effect of space charge on the period of the electron motion. This period is one half that of the sine function in (11), since the electron leaves the cathode, approaches the plate, and returns to the cathode, twice during a single period of (11). Hence the electron period τ is given by

$$\tau = \frac{4\pi m}{(2-n)He}. \tag{12}$$

It is evident that space charge may radically affect τ, since as n approaches 2, τ approaches infinity.

Several special cases are of interest. If $n=0$, $V=V_p$ and the electron receives all its acceleration at the start, and continues to move henceforth with uniform linear velocity. According to (12) the period is $2\pi m/He$, which is the familiar formula for an electron of constant linear velocity in a uniform magnetic field. This case gives the highest possible frequency unless V at some point exceeds V_p. The value of the product λH, which should be a constant, according to Okabe,[8] is in this case 10,700.

For the case of space-charge limitation, in the absence of oscillations and a magnetic field, $n=2/3$, and $\tau=3\pi m/He$. Hence $\lambda H=16{,}057$. A value of 16,700 was obtained for this case by Megaw[9] using a graphical integration method.

[8] K. Okabe, Proc. I.R.E., vol. 17, pp. 652–659; April, (1929).
[9] E. C. S. Megaw, Proc. I.R.E., vol. 21, pp. 1749–1751; December, (1933).

Values of n greater than two thirds do not occur for space-charge limitation in the absence of oscillations and a magnetic field except near the surface of a filament of finite size where $n=4/3$. However, very little is known of the space-charge distribution during oscillation, and there are no grounds for ruling out of consideration larger values of n.

In case $n=2$, $\tau=\infty$. The equation of motion (9) takes the simple form

$$\frac{dr}{dt}=\frac{r}{r_p}\sqrt{\frac{2eV_p}{mr_p^2}-\left(\frac{He}{2m}\right)^2}=r\text{ const.}$$

If $V_p>H^2er_p^2/8m$, the constant is real, the velocity is proportional to r, and the radial motion is aperiodic. If $V_p=H^2er_p^2/8m$, the constant is zero, and the radial velocity is likewise zero. In this case it can be shown that the radial electric force exactly balances the centripetal force so that an electron once started in circular motion about the filament as center, with radius r, and velocity corresponding to $V(r)$, will continue to move thus indefinitely. If $V_p<H^2er_p^2/8m$, the velocity is imaginary.

If $n>2$, the radial motion again becomes periodic, but as pointed out above, does not satisfy the condition, $r=0$, when $t=0$.

Further information on the effect of space charge can be obtained by considering the variation of r_1, the maximum radial excursion of the electron from the cathode. This is obtained from (11) by giving the sine factor its maximum value, unity. This yields

$$\frac{r_1}{r_p}=\left(\frac{8mV_p}{H^2er_p^2}\right)^{1/(2-n)},\tag{8}$$

agreeing with the expression previously found in the section on static characteristics.

Three cases need be considered,

$$\left.\begin{array}{lll}(a) & V_p = & \dfrac{H^2er_p^2}{8m}\\[2ex] (b) & V_p < & \dfrac{H^2er_p^2}{8m}\\[2ex] (c) & V_p > & \dfrac{H^2er_p^2}{8m}\end{array}\right\}.\tag{13}$$

The first of these has been discussed by Hull.[2] He shows that in this instance $r_1=r_p$, and that this is true regardless of the form of the

potential function $V(r)$. This same result follows readily from (8), since if (13a) be substituted, the quantity in the parenthesis becomes unity and $r_1 = r_p$ for all values of n. This case is represented by the horizontal dashed line a in Fig. 16.

If $V_p < H^2 e r_p^2/8m$, equation (8) gives the curve of Fig. 16(b). The dash-dotted section corresponding to $n > 2$, need not be considered, since, as was shown above, $r \neq 0$ when $t = 0$. In the region $n < 2$, the curve shows that r_1 decreases as space charge increases, and approaches zero as n approaches 2.

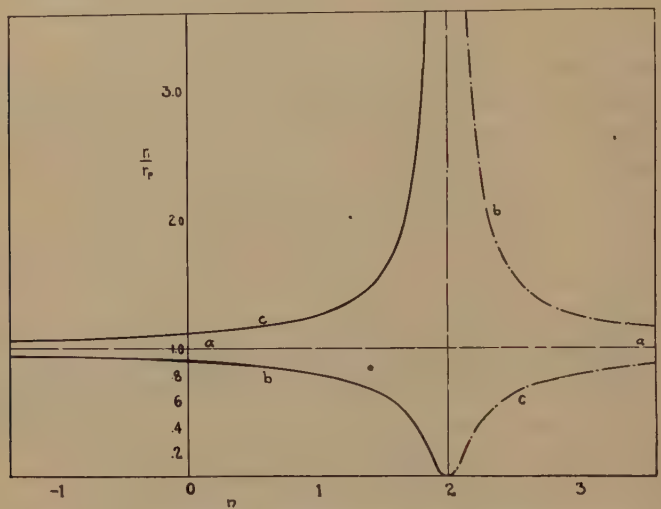


Fig. 16—Maximum radial excursion of an electron as a function of n in equation (7).

The third case, $V_p > H^2 e r_p^2/8m$, is shown in Fig. 16(c). These curves are the reciprocals of curves b. As before, only the region $n < 2$ need be considered. Here the variation of r_1 is in a direction opposite to that of the previous case, r_1 increasing as space charge becomes greater. Of course, the electrons would strike the plate before r reached its maximum value r_1.

From Fig. 16 it may be concluded also that the sensitiveness of r_1 to changes of plate voltage increases as n approaches 2. This follows from the increasing separation of curves (b) and (c). The curves in the figure are drawn for the cases,

$$V_p = 0.8 \frac{H_c^2 e r_p^2}{8m} \text{ for curve } b, \text{ and}$$

$$V_p = 1.25 \frac{H_c^2 e r_p^2}{8m} \text{ for curve } c.$$

Experimental evidence of the variation of r_1 with n is given in Fig. 17. Here I_s, the temperature limited filament emission in the absence of a magnetic field, is plotted against I_p, the plate current with the magnetic field applied. Increasing I_s is equivalent to increasing n, particularly so when the magnetic field has the cutoff value so that the electrons make many revolutions before striking the plate, and thus augment the space-charge effect.

In obtaining Fig. 17, a tube without end plates was used. The magnetic field was adjusted to a value near the upper bend of the

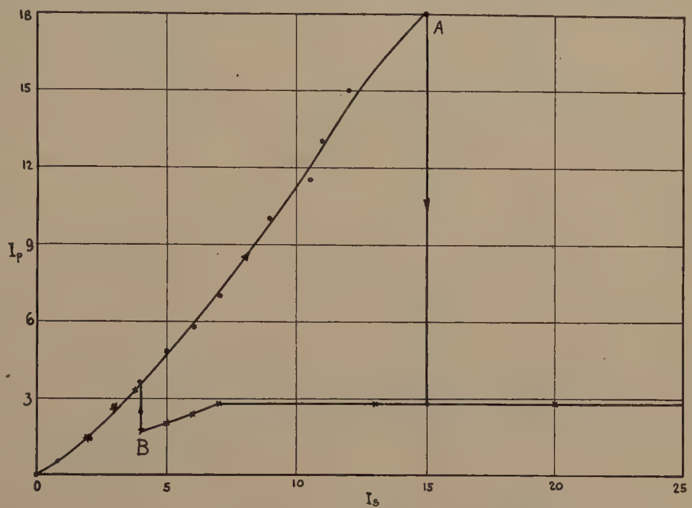


Fig. 17—Effect of space charge on cutoff of plate current. Plate current is plotted against the temperature limited filament emission to be expected in the absence of a magnetic field.

cutoff curve (see Fig. 6). Upon increasing I_s it is found that I_p increases almost linearly at first, but eventually a point A is reached at which I_p suddenly drops to a much smaller value. This is explained as caused by the decrease of r_1 due to the increase of n, which results in more complete cutoff of the plate current.

The abruptness of the drop is greater than would be expected for such a relation as (8). This is apparently due to two causes. First, there is a cumulative action, for when r_1 begins to decrease, cutoff becomes more complete, the space-charge effect is augmented, and r_1 is further decreased. This increase is in addition to that directly caused by I_s. Second, as cutoff becomes more complete, electron bombardment of the filament begins to occur (see section below on instability), raising its temperature so that the actual filament emission exceeds I_s. This effect builds up very rapidly as the critical value of I_s is approached.

Both of the above effects produce an unstable state, which results in an abrupt rather than a smooth drop in plate current.

Beyond the point A, the plate current I_p is space charge limited. The low value of current at which this occurs is due to the long helical paths traversed by the electrons in traveling from cathode to plate. By tilting the tube these paths are shortened, and the space-charge-limited current can thus be increased.

Upon decreasing I_s an upward jump does not occur at A, but at B, corresponding to a lower I_s. The return curve points are indicated by

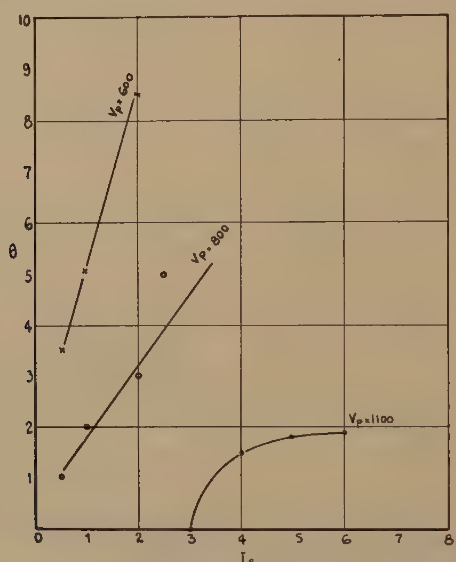


Fig. 18—Relations between optimum angle of tilt and temperature limited filament emission for various plate voltages.

crosses. The occurrence of the jump B at a lower I_s than A is apparently due to heating of the filament by electron bombardment so that the true emission is greater than the I_s plotted.

The decrease of r_1 with increase of n can be revealed also by studying the variation of θ_{opt} with I_s. As the latter quantity increases, r_1 decreases and hence a greater value of θ should be required to eliminate a sufficient number of electrons to yield the optimum space-charge condition, providing V_p be maintained constant. This is found to be the case, as shown by the data of Fig. 18. Here θ_{opt} is plotted against I_s for various values of V_p. For the smaller values of V_p the relation seems almost linear, θ increasing with I_s, as is to be expected. As V_p is increased, r_1 increases and hence less tilt is required. If V_p is close to the value at

which r_1 is very nearly equal to r_p, and at which the plate can eliminate sufficient electrons without tilt, the relation is no longer linear, as shown for $V_p=1100$. The closer V_p is to the value for which $r_1=r_p$ the less does θ_{opt} vary with I_s.

When end plates are used, θ_{opt} does not vary with I_s. Substantially the same proportion of electrons are removed for all values of I_s. The optimum condition of operation of the end-plate magnetron is, therefore, not greatly dependent on either V_p (see Fig. 13) or I_s, and therein it possesses a marked advantage over the simple type.

The above discussion and data have shown that the effects of space charge are of prime importance, and the alterations of path dimensions (r_1), frequency, and sensitiveness to plate-voltage changes, are amply sufficient to make the space-charge condition a critical factor. The existence of an optimum space-charge state, as was suggested by the experimental data, is easily understood in view of the above discussion.

Limitation of Output by Instability

In common with other types of electronic oscillators, the end-plate magnetron is subject to a peculiar type of instability, which at present is the chief obstacle towards obtaining greater output. This instability appears as a sudden brightening of the filament and simultaneous increase of tube currents, when attempts are made to increase the oscillation energy.

A study of this phenomenon carried out by the writer indicates that it arises from electron bombardment of the filament. Beyond the cutoff point there has been shown to exist in the tube a space charge of electrons having energies greater than that corresponding to the direct-current potential they have fallen through. They therefore return to the filament with velocities greater than zero, and by bomardment increase its temperature. An article discussing this instability will appear later.

The data presented with this paper were obtained under conditions of low energy oscillation. The phenomena associated with instability therefore are not involved, except when otherwise stated. With more energetic oscillations, instability radically changes the form of many of the curves, and obscures the fundamental behavior.

Acknowledgment

The writer is indebted to his associates for very helpful co-operation, and especially to Dr. Irving Wolff for valuable discussions and advice.

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SUPPLEMENTARY NOTES ON "ANALYSIS OF THE OPERATION OF VACUUM TUBES AS CLASS C AMPLIFIERS"*

By

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During a discussion of the above paper after its presentation at the tenth annual convention at Detroit, questions were asked concerning the exact location of a dynamic characteristic satisfying the condition of "double equilibrium." Is it to be found in each case by pure trial and error, or can it be arrived at in some systematic way?

Undoubtedly, an engineer after having practiced the described method on a few samples will always be in a position to predict with sufficient accuracy where and how the appropriate characteristic will be located. However, if mathematical accuracy is desired, the exact location of a straight-line oscillation path at any plate voltage during modulation can be established by the following systematic procedure.

Suppose the carrier condition at plate voltage E_c has been chosen by the method given in the paper (or by any other method) so that the load resistance, R_L, the grid-leak resistance, R_g, and the grid excitation amplitude, e_0, are definitely fixed. Now it is desired to find the exact location and all operating data for any other plate voltage during the process of modulation, for example, at its positive peak, equal to $2E_c$. For this plate voltage we choose several different operating points corresponding to several assumed biases. Through each point we plot a straight-line dynamic characteristic with the vertical projection equal to the amplitude of the grid excitation voltage (established by the carrier condition) and having for its horizontal projection an arbitrarily assumed radio-frequency plate voltage swing E_0. In other words we also choose arbitrarily a minimum plate voltage, $E_{\min} = E_p - E_0$, which will be touched during oscillation. Such characteristics will be of course parallel to each other. The average grid current calculated along each line multiplied by R_g will give us the true bias for the assumed position of the dynamic line; we can also calculate for each line the true power output from the tube. Generally speaking, none of the plotted characteristics will satisfy either of the conditions of

* Decimal classification: R132. Original manuscript received by the Institute, August 29, 1935. See Proc. I.R.E., vol. 23, pp. 752–778; July, (1935), for original paper.

double equilibrium. The calculated biases will be different from those assumed in plotting each individual dynamic characteristic, and power output calculated for each path will not be balanced by the power consumed in the load resistance, R_L. However, by interpolation we can find where a dynamic characteristic should be located in order that the condition of either output power, or of grid-bias equilibrium be satisfied. Generally, these will be two *different* dynamic lines, witnessing that—with the assumed minimum plate voltage—there is no possibility of a stable operating condition.

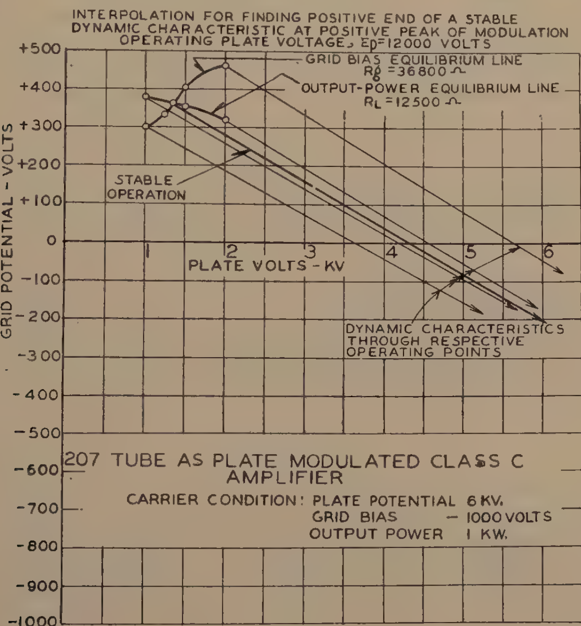

Fig. 1

A similar procedure is to be repeated for several other minimum plate voltages. In this way, we shall be able to trace on the chart two curves connecting the ends of the dynamic lines satisfying one, or the other equilibrium condition at different minimum plate voltages. Naturally, their intersection gives the only point of perfectly stable operation. With a little experience one can conveniently limit the calculation to three or even two dynamic lines for each assumed minimum plate voltage and, also to three or two different minimum plate voltages.

The described graphical procedure is illustrated in the accompanying figure which has been drawn for a special case of operation of the

207 tube indicated in the drawing and which, in the light of the above discussion, is self-explanatory.

It is perhaps interesting to point out in this place that from the educational view point, the "constant-current charts," described in the paper, may be of great help in explaining to a student various phenomena in self-oscillators, such as the rôle of grid coupling to the oscillating circuit, the mutual relation between the grid coupling and the amplification factor of the tube, and the relative ease of starting oscillations at various conditions; in short, the entire mechanism of oscillation of a tube.

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A NOTE ON THE GRAPHICAL ANALYSIS OF ALTERNATING-CURRENT NETWORKS*

By

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Summary—*A simple proof is given of the following observed property of alternating-current networks: In a network consisting of any number of linear bilateral impedances, connected in any manner whatsoever, and any number of generators of the same frequency, the locus of any voltage or current in the system will be a circle as any one self-impedance is varied so that its locus is a straight line in the complex plane. The proof given is based upon the application of the fundamental theorems of network analysis to a general alternating-current network and upon the identification of a circular form of the linear fractional transformation of function theory. A simple analysis of alternating-current networks is frequently possible by virtue of this property. Reference is made to examples of circuit analysis based upon this property. It is further shown that as a result of the observed property the impedance looking into any two terminals of a general passive network of linear bilateral elements has a circular locus as any impedance in the system is varied so that its locus is a straight line.*

WITHIN recent years a number of papers have appeared dealing with the graphical analysis of alternating-current networks. Several of these[3,4,5]† have included analyses based upon the use of a form of the circle

$$C = \frac{Lx + M}{Px + Q}, \tag{1}$$

the linear fractional transformation of complex function theory which gives C as a circle in the complex plane when x is taken as a real number varying from minus to plus infinity, and L, M, P, and Q are any complex numbers.

The papers referred to above have treated specific cases of the circular form of current, voltage, and impedance for particular types of impedance variation. They have given analytical expressions showing the loci in question to be circular, and have given means enabling their location and the rapid determination of the values desired from them.

A simple proof is given here of the property of alternating-current networks which makes the above type of analysis possible. This property for the case of a general network may be stated as follows: In a network consisting of any number of linear bilateral impedances, connected in any manner whatsoever, and any number of generators

* Decimal classification: R140. Original manuscript received by the Institute, August 30, 1935.

† Numbers refer to Bibliography.

of the same frequency, the locus of any voltage or current in the system will be a circle as any one self-impedance is varied so that its locus is a straight line in the complex plane. It follows from this that the locus of the impedance at any two terminals of a general network is also a circle under the same conditions of variation.

Use will be made of several of the relations of complex function theory:

That a straight line in the complex plane can be represented by

$$S = Px + Q \tag{2}$$

where P and Q are any complex numbers and x is a scalar variable;

That the reciprocal of a straight line in the complex plane, a circle through the origin, may be represented by

$$C_0 = \frac{1}{S} = \frac{1}{Px + Q}; \tag{3}$$

That a general circle in the complex plane may be regarded as the sum of a constant and a circle through the origin, the expression for which assumes the form of the quotient of two straight lines which is the linear fractional transformation form referred to above. Symbolically, the general circle is given by

$$C = K + \frac{1}{S} = K + \frac{1}{Px + Q} \tag{4}$$

$$= \frac{Lx + M}{Px + Q} = \frac{S_1}{S_2} \tag{5}$$

where $L = PK$ and $M = QK - 1$.

The demonstration of the property stated above falls into two parts; first, consideration of the form of the current through and the voltage across the impedance being varied, and second, consideration of the form of the current through and the voltage across any other impedance in the system.

The proof of the circular form resulting in the first case is almost superfluous, involving as it does no more than the interpretation of Thevenin's theorem in terms of loci in the complex plane.

Let Z be the impedance of the network element being varied. All other impedances of the system and all electromotive forces are held constant. A variation of Z so that its locus is a straight line will usually mean either a variation in magnitude at a constant phase angle or a variation along its resistive or reactive component. Most of the well-known circle diagrams can be reduced to an analysis based on a linear impedance variation like this.

By Thevenin's theorem, the network of impedances and generators feeding the impedance, Z, can be replaced by a simple generator whose generated voltage, E_{12}, is the open-circuited voltage at the terminals of Z and whose impedance, Z_i, is the impedance of the network looking back from the terminals of Z with all generators replaced by their internal impedances.

The current, I_z, through the impedance Z is given by

$$I_z = E_{12}/(Z + Z_i). \tag{6}$$

Speaking in terms of loci, Z is a straight line, if this impedance is so varied, and Z_i is a constant vector. Therefore the denominator on the right side of (6) is a straight line. I_z is thus seen to be a constant, E_{12}, times the reciprocal of a straight line giving a circle through the origin. Thus, the locus of the current through an impedance as this impedance is varied so that its locus is a straight line is a circle through the origin.

Consider the voltage, V_z, across the impedance now being varied in the first case. Using the value of I_z from (6),

$$V_z = I_z Z = \frac{E_{12}Z}{Z + Z_i}. \tag{7}$$

The numerator of the fraction in (7) is a constant vector, E_{12}, times a straight line, Z, and thus a straight line. The denominator is a straight line as above. Therefore V_z appears as the quotient of two straight lines, and hence its locus is a general circle by (5).

For the second case, consider the current through any impedance other than that which is being varied. The conditions obtaining are shown in the diagrammatic representation of part (a) of the schematic circuit of Fig. 1. Let Z, the impedance being given a straight line variation, be divided into a constant component, Z_c, and a variable component, Z', with Z_c so chosen that the locus of Z' passes through the origin. A separation of this sort, in fact an infinite number of them, can always be effected. This separation may be represented by

$$Z = Z_c + Z'. \tag{8}$$

From the vector diagram of Fig. 2 showing this relation, it is seen that Z' will be a straight line through the origin parallel to the straight line locus of Z.

Let Z_w be any impedance of the system other than the varied impedance and I_w the current through it. By the principle of superposition I_w can be considered as being made up of a constant component of current, I_c, which flows when $Z'=0$ and $Z=Z_c$, plus a vari-

able component of current, I_w', which is due to the voltage $V_z' = -I_z Z'$. Thus

$$I_w = I_c + I_w'. \tag{9}$$

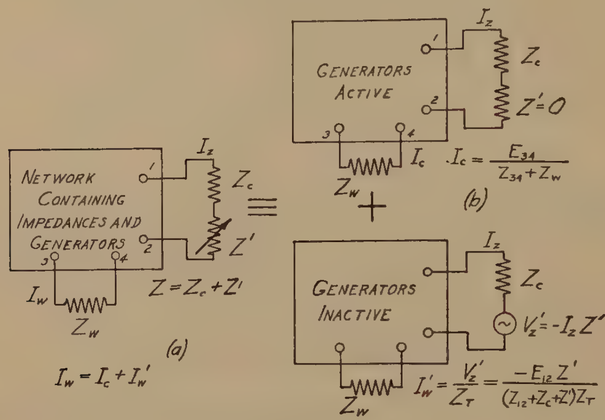

Fig. 1

The schematic representation of Fig. 1(b) shows the circuit broken to produce these components of current. In a circuit of linear bilateral

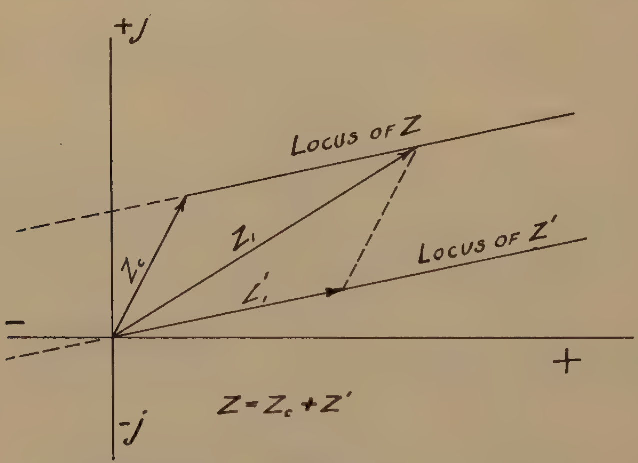

Fig. 2

impedances, I_w' will be a linear function of V_z' and will be given by

$$I_w' = V_z'/Z_t \tag{10}$$

where Z_t is the transfer impedance connecting the two. Substituting the equivalents of V_z', I_z, and Z in (10)

$$I_w' = \frac{-I_z Z'}{Z_t} \tag{11}$$

$$= \frac{-E_{12}}{Z_i + Z} \frac{Z'}{Z_t} \tag{12}$$

$$= \frac{-E_{12}}{(Z_i + Z_c) + Z'} \frac{Z'}{Z_t}. \tag{13}$$

In terms of loci, the numerator of the right side of (13) is a constant, $-E_{12}$, times a straight line through the origin, Z', and thus a straight line through the origin. The denominator is the product of a constant vector, Z_t, and a straight line $(Z_i+Z_c)+Z'$, and thus a general straight line in the complex plane. Hence I_w' is the quotient of a straight line through the origin and a general straight line and is thus a circle through the origin. Substituting the value of I_w' from (13) into (9)

$$I_w = I_c + \frac{-E_{12}}{(Z_i + Z_c) + Z'} \frac{Z'}{Z_t}. \tag{14}$$

From this I_w represents the difference of a constant vector, I_c, and a circle through the origin, I_w', and is thus a general circle in the current plane. Thus as any self-impedance in the system is varied so that its locus is a straight line, the locus of the current through any other impedance in the system is a circle.

The voltage across any impedance Z_w other than that being varied is given by

$$V_w = I_w Z_w. \tag{15}$$

Since I_w is a circle by (14), multiplication by a constant vector, Z_w, will only change the scale and rotate the figure leaving the locus still a circle. Thus as any self-impedance in the system is varied so that its locus is a straight line, the locus of the voltage across any other impedance in the system is a circle. The property is thus proved for all cases.

The proof given above, though by no means the simplest possible,* has attempted to keep in mind the physical picture and to show intimately the relation between the individual elements involved.

From the above property it can be shown that the impedance into any two terminals of a general passive network of linear bilateral impedances will have a circular locus as any impedance in the system is

* In a letter to the author, Hans Roder points out that the proof follows immediately from writing down the expression for any current or voltage in determinant form.

varied so that its locus is a straight line. If a generator with a constant electromotive force, E, and an internal impedance Z_g is connected across the terminals of a general network, the current

$$I = E/(Z_g + Z_i) \tag{16}$$

which will flow will be a circle as any impedance in the system is given a straight line variation, by the property discussed above. Solving for the impedance into the terminals

$$Z_i = E/I - Z_g. \tag{17}$$

In terms of loci, Z_i is the difference between the product of a constant vector, E, and the reciprocal of a circle, $1/I$, and another constant vector, Z_g. The resulting locus is a circle since the reciprocal of a circle is a circle and rotation, change of scale, and translation do not change the circular form.

The circular loci of cross potential and cross current of a bridge circuit operating at a constant frequency as given by Seletzky[3,4] are examples of the general property discussed here. Other more common examples are the induction motor current circle as based upon the equivalent transformer circuit and the air-core transformer current and voltage diagrams. The circular form of the input impedance of a radio frequency transmission line with varying load impedance as given by Roder[5] is an example of the circular nature of impedance loci.

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BOOK REVIEWS

"The Radio Engineering Handbook," by Keith Henney. Published by the McGraw-Hill Book Company. Price $5.00. 831 pages and an 18 page Index.

The editor of a handbook is confronted with the difficult problem of including most of the information for which the engineer has frequent need and at the same time keeping the size of the book within reasonable limits to facilitate its use. In the second edition of the Radio Engineering Handbook, Henney has attempted to meet the first requirement by adding 300 pages of new material, an increase of about fifty per cent over that of the first edition. The size has not been unduly increased although a good grade of heavy paper is used. A corresponding increase of new material in another edition will require the use of onion skin paper or the elimination of some of the less important information now included.

Probably the most significant features are the expansion of the section on radio broadcasting from twenty to sixty-three pages and the inclusion of a new sixty-eight page section on antennas. Certainly these will be appreciated by many users. The continued progress in the several phases of radio engineering are reflected in new data and new information included. The authors, for the most part, are well known with established reputations which provide the authoritative element. In the case of a few sections there have been changes in the authors.

The following are a few illustrative criticisms. The most apparent deficiency, from my point of view, is the lack of hyperbolic tables. I object to the form of the tabulation on pages 6, 7, 8, and 9 for they are incorrect without the multiplying factors given on page 5 to which there is no reference. The tabulation should be correct for the frequencies given without the multiplying factor. The double heading for the second column is misleading for obviously it cannot be true for both ω and ωL unless $L=1$ in which case there would be no point in including it. The same criticism applies to the double heading of column three. One would scarcely expect to find under the heading "General Characteristics of Alternating-Current Circuits" the equations for the transient currents with a constant voltage impressed. In the discussion of the ratio of high-frequency resistance to direct-current resistance no mention is made of whether the wire is a straight conductor or wound into a coil of many turns. On page 73 the induced electromotive force is given without restriction as

$$e = -\frac{d\phi}{dt} \text{ instead of } n\frac{d\phi}{dt}.$$

*H. M. Turner

"The Radio Amateur's Handbook," Published by American Radio Relay League, Hartford, Conn. Price $1.00. 380 pages.

To those who are familiar with previous editions of this manual, it is only necessary to state that the latest edition is comprehensively revised in basic

* Yale University, New Haven, Connecticut.

material and includes much new subject matter. All material covering circuits and transmitter and receiver construction is brought thoroughly up to date.

It is not possible here to outline in detail all of the material of the 380 pages of text, which is divided into twenty-one chapters. Two chapters deal with ARRL history and introducing the subject of amateur radio to those just entering this activity. Four chapters are devoted to electrical principles, circuit and wave fundamentals, vacuum tubes, and fundamentals of radiotelephony. Tube data tables include late types of both metal and glass receiving tubes, as well as special purpose, rectifier and transmitting tubes.

Six chapters are devoted to receiver and transmitter design and construction including equipment for the ultra-high frequencies, for both code and voice communication. In these chapters, the segregation of this material into "design" and "construction" groups makes for clear presentation of design considerations without fogging due to constructional details.

Chapters on keying, covering various systems, methods for key-click elimination, and for remote control, power supplies, antennas and instruments and measurements complete the technical material. All are up to date and give full information for construction and operating adjustment.

The final chapters cover installation (of a station as a whole), operating procedure, message handling and ARRL operating organization.

There are but few unfavorable features, and these are of such a nature as to no way impair the usefulness of the Handbook to those for whom it is intended. Irregularities in style and presentation of material are probably unavoidable in a compilation of this size by many individuals. The complete absence of references, either to *QST* or other sources, is regrettable as it detracts greatly from the value of the work as a contemporary account of progress in amateur radio.

While intended for amateurs and near amateurs, this Handbook should be of interest and use to radio engineers and commercial operators as well.

†James K. Clapp

† General Radio Company, Cambridge, Mass.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed below may be obtained without charge by addressing the publishers. They are not available from the Institute.

"The Collins Signal" is published at irregular intervals by the Collins Radio Company of Cedar Rapids, Iowa.

Yaxley Replacement Manual and Service Guide No. 2 is available from the Yaxley Manufacturing Division of P. R. Mallory and Company, Inc., Indianapolis, Ind.

Amateur radio equipment is described in catalog No. 60 issued by Wholesale Radio Service Company of 100 Sixth Avenue, New York City.

Bliley quartz crystals are described in a catalog under that name issued by the Bliley Electric Company of Erie, Pa.

"The Electronic Parade" is the name of a catalog issued by Electronic Laboratories, Inc., of Indianapolis, Ind., and covers vibrators, testers, converters, and buffer condensers.

A number of products developed by the Brush Development Company, East 40th St. and Perkins Avenue, Cleveland, Ohio, are illustrated in a leaflet issued by that organization.

Supreme Instruments Corporation, Greenwood, Miss., has issued a 1936 catalog which covers their receiver test equipment.

The Ken-Rad Corporation of Owensboro, Ky., has issued engineering bulletin CEB 36-5 entitled "The Relation of Modulation Products with Multi-Tone Signal to Harmonic Distortion with Mono-Tone Signal in Audio Amplifier Analysis."

National Union Laboratories, 365 Ogden St. Newark, N.J., has released an engineering bulletin on 6R7MG duodiode medium-mu triode.

The Hygrade Sylvania Corporation of Emporium, Pa., has issued engineering news letter No. 19 on audio output systems for battery-operated receivers, No. 20 on improving diode detector performance, and No. 21 on characteristics of audio output systems using type 1F4 tubes. Technical data sheets have been issued on type 1F4 power output pentode and the 6R7 duodiode medium-mu triode.

RCA Manufacturing Company of Harrison, N.J., has issued application note No. 55 on the operation of the 6A8, No. 56 on receiver design, and No. 57 on the 6L7 as a radio-frequency amplifier.

Shure Brothers, 215 West Huron St., Chicago, have issued leaflets describing their "Spheroid" microphone and their microphone repair service.

The Ward Leonard Electric Company of Mt. Vernon, N.Y., has issued bulletin No. 1105 on ring type rheostats, 5701 on a constant voltage variable load regulator for alternating current circuits, 8602 on controlled rectifiers to supply direct-current power for small telephone systems, and 8603 on a controlled rectifier to supply small direct voltages from an alternating-current power supply.

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